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Design Techniques for Interference-Free Operation of Airborne Electronic Equipment - and Appendixes I thru XVII

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TO: Recipients of Handbooks

This Handbook on Design Techniques for Interference-Free Operation of Airborne Electronic Equipment has been prepared to further promote the art of interference suppression on military equipments.

It is primarily intended for the design engineer of electrical and electronic equipment. Recognition of the interference problems and mutual interaction between electronic systems at the time of the initial design will alleviate most of the interference difficulties encountered in the operation of military aircraft.

It is intended to revise this publication when enough additional data becomes available to warrant a revision. It is requested that comments and additional interference suppression design data which would contribute to the purpose of this Handbook be forwarded to the following address: Commanding General, Wright Air Development Center, Wright-Patterson Air Force Base, Ohio, Attention: WCESI-5.

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FOR THE COMMANDING GENERAL:

Carl P. Walter

CARL P. WALTER
Colonel, USAF
Chief, Components and
Systems Laboratory
Weapons Components Division

DESIGN TECHNIQUES

For

INTERFERENCE-FREE OPERATION

of

AIRBORNE ELECTRONIC EQUIPMENT

Prepared By

FREDERICK RESEARCH CORPORATION

Bethesda, Maryland

on

CONTRACT AF 33(038)23341

1952

UNITED STATES AIR FORCE

WRIGHT AIR DEVELOPMENT CENTER

WRIGHT-PATTERSON AIR FORCE BASE

DAYTON, OHIO

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ERRATA

Page 1 - 30, line 1, correct to read:
$L < r$

Page 4 - 7, line 15, correct to read:
Paragraph 4. 2. 4. 1

ABSTRACT

Presented are design techniques for interference-free operation of airborne electronic equipment which may be used as a guide by the aircraft and equipment designer. Section I contains a discussion of the basic theory of interference which is applied to illustrative components and system problems in Section III. Section II briefly discusses several applicable interference specifications together with methods of measurement and acceptable interference test sets. Section IV is devoted to precipitation static and techniques for its suppression.

FOREWORD

This book is a treatise on existing radio interference problems and solutions. It shows how the known solutions are rooted in theory. More than that, it is, as its title denotes, a guide for all designers of aircraft and manufacturers of airborne electronic equipment. Its purpose is to present a comprehensive treatment of successful techniques of interference suppression and to start the design engineer thinking early - before his functional design is finished and he winds up with a "noisy" equipment which is of no value to the Air Force.

Because of the rapid advance in radio and allied fields, the authors are fully aware that certain sections of the book may eventually have to be interpreted in the light of newer designs and equipments. The first section of the book is, therefore, very basic in hopes that a clear presentation of the theory of interference will be of great value in solving the inevitable problems which result from the introduction of new electronic equipment and new designs.

The research and preparation of manuscript as well as the printing of this book was accomplished by the Frederick Research Corporation under Contract AF33(038)23341. Work was initiated by, and accomplished under the direction of, the Radio Interference Unit, Components and Systems Laboratory, Weapons Components Division, Wright Air Development Center, Mr. N. D. Flinn acting as project engineer.

By furnishing the material in this book in condensed and useable form, together with the appendices, list of selected references and index, it is hoped that a significant contribution has been made toward the design and production of interference-free aircraft equipment.

INTRODUCTION

The complexity of the problem of radio interference has increased tremendously, in recent years, owing to the large increase in electronic equipment installed in operational aircraft and has become one of the most serious handicaps to the operation of the United States Air Force.

This problem has been attacked from many different angles by the Armed Forces, aircraft designers, and by electronic equipment manufacturers. Electronic Engineers - both military and civilian - have made rapid strides in combating this deterrent to operational efficiency and, as a result, many papers on the subject have appeared in technical journals. Also several technical manuals have been prepared by each of the Military Services. All these have been helpful and have emphasized the pressing need for an up-to-date book, comprehensive in its treatment of the subject.

Radio Interference greatly decreases the Reliability and Efficiency of Operational Aircraft because of the loss of intelligibility of vital information or the loss of accuracy of interpretation by Receivers whether presented acoustically, as in earphones, visually on cathode ray scopes, or mechanically as in the servo operation of instruments and aircraft controls.

Basically, Interference Reduction Techniques are an application of electrical transmission theory and good electromechanical engineering practice. An effort has been made in preparing this book to bring out the basic principles involved in solving radio interference problems and point out the best known practices for use of the designer. In some cases poor design practices are also discussed in order to emphasize the practices to avoid and also to impress the reader with the waste involved in having to apply "Fixes" in the field.

Interference may be classified according to the way in which it is generated as follows:

1. Nature-made. This interference is caused by atmospheric electrical disturbances and precipitation static.

2. Inherent. This is interference generated within the receiver due to thermal agitation, shot effect, and similar causes.

3. Man-made. This includes interference from sources both within (e.g., motors) and without (e.g., jamming) the aircraft.

Interference may also be classified according to the place where it is generated as follows:

1. Interference generated within, on, or in the immediate vicinity of the aircraft.

2. Interference generated outside and entirely independently of the aircraft.

INTRODUCTION

Because this book is intended primarily for the design engineer and must be kept within reasonable size for convenient reference, it deals only with a narrow selection of interference phenomena. The book, therefore, begins with a treatment of the basic theory of interference and how it is generated within the aircraft in order to help the design engineer to apply the fundamental principles to his specific problem if not covered in the application given in Section III. The section on measurements briefly reviews some of the pertinent government specifications which the designers should constantly review owing to the rapid advances being made in the art of interference measurements. A section on Precipitation Static is included because of its importance in the flight of operational aircraft, and it's effect on the functional operation of airborne electronic equipment.

The criticism has often been made that measurements and tests for radio interference are too often made under ideal conditions without any anticipation of the many adverse conditions existing in actual flight, which fact may wholly or partially nullify the validity of these tests - theoretically and practically. It has been seen, too, that individual laboratories have conscientiously attempted to eliminate or forestall radio interference within their own equipment but in such a way that there is little or no appreciation of the interaction of the various equipments within the aircraft. As a result, many radio interference problems are far from being solved. Successful solutions require team work all along the line including the original designer and the installation and maintenance personnel.

The very remedies instigated for certain specific interference problems have, in some cases, actually made the "noise" worse and constituted new interference problems in other equipment. This often results from "fixes" being undertaken without a thorough study of the basic theory underlying the attempted fix.

For the purposes of this book, RADIO INTERFERENCE is defined as any electrical disturbance which causes an undesirable response or malfunctioning in any electronic equipment. Any audible, visible, or otherwise measurable response is considered undesirable if it is not produced by a desired signal, provided that either its duration is longer than one second or its highest recurrence rate during normal operation of the aircraft is greater than once every three minutes.

The word "receiver" will be used in a generalized sense to include all electronic equipment in which interference may cause undesirable response or malfunctioning. For purposes of this book, therefore, a RECEIVER is defined as any electronic equipment in which unwanted signals may cause an undesired response.

* * * * * * *

Each section and appendix of this book has an independent sequence of page numbers. For example, 3 - 15 indicates page 15 of Section 3 and V - 3 indicates page 3 of Appendix V.

* * * * * * *

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SECTION I - THEORY

1 THEORY OF INTERFERING SIGNALS

The first step in combating an evil must always be an attempt to gain a thorough understanding of its nature. To this end, a detailed theoretical analysis of many of the problems and phenomena associated with radio interference will be presented in this section.

For the purposes of this book, radio interference is defined as any electrical disturbance which causes an undesirable response or a malfunctioning in any airborne electronic equipment. The word "interference" was chosen in preference to the word "noise" in order to emphasize the fact that the disturbing signals to be treated may lie in the frequency range from zero up to the highest used in existing or proposed aircraft. It is true that the term "radio noise" is commonly used to denote undesired signals of any frequency. Yet, strictly speaking, "noise" is an acoustical disturbance, confined to frequencies below approximately 20 kilocycles per second, and this connotation may easily create the wrong impression in the reader's mind.

There is still another reason which makes it advisable to use the word "interference" instead of the word "noise" in this book. According to the above definition, a signal is not to be considered as interference unless it causes an undesirable response or false operation. In most cases, the output of the equipment which is affected by the disturbance is an audio signal, and the term "noise" properly applies to this interference in the output. The definition of the word "noise" may be extended to include, also, the disturbances in the output of those devices which give visual indications of one kind or another. But any further extension of the definition of this term is likely to lead to confusion because it may cause one to forget that interference may be of any frequency even though its final product, the disturbance in the output, is usually in the audio range.

It should be mentioned here that, for brevity, the definition of the word "receiver" is extended in this book to include all types of electronic equipment in which interference may cause an undesirable response or malfunctioning. Thus, not only an earphone, but a navigation instrument or a radar scope, will be called a receiver, and even a relay, which may be tripped by a spurious signal, will fall under this heading.

1.1 STATISTICAL ANALYSIS

In statistical theory, the term "random noise" is used for disturbances that are completely without regularity in their detailed properties. These disturbances lend themselves well to treatment by statistical methods, and their theory has been extensively discussed in the literature. An example of interference which can be so treated is the tube noise due to thermal agitation and the shot effect. Into this group also belong atmospheric disturbances commonly known as "static". The characteristics common to these phenomena are their random amplitudes, random phases, and lack of periodicity. If a frequency analysis is performed on such disturbances,

it is found that their energy is fairly uniformly spread throughout the frequency spectrum from zero up to a maximum (sometimes as high as 10 kilo-megacycles per second, depending upon the source) and drops to zero rather rapidly for frequencies above that maximum. A plot of energy versus frequency for such disturbances will, therefore, have a shape somewhat as shown in Figure 1.1-A. This "random noise" consists of an irregular sequence of pulses of arbitrary shape which bear no relation to each other. Most of the interference to be treated in this book does have some regularity. It is what in statistical theory would be classified as a "signal" rather than "random noise". The difference lies in that these interfering signals are usually periodic and show some regularity although their wave shapes and phases may be subject to statistical fluctuations. Like "random noise", their energy is often spread over a very wide frequency spectrum. Unlike "random noise", the distribution of their energy is usually non-uniform, showing wide variations with definite maxima and minima as for example in Figure 1.1-B.

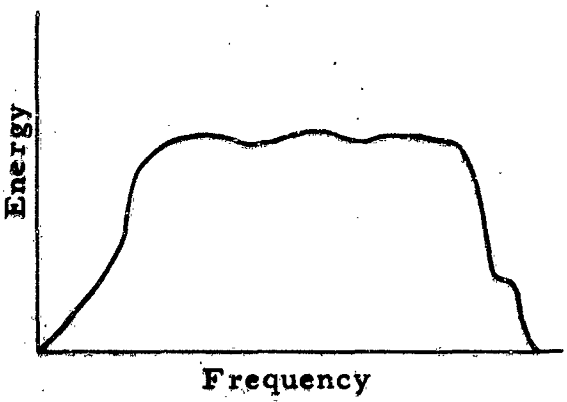

Fig. 1.1-A Typical Energy Distribution of "Random Noise"

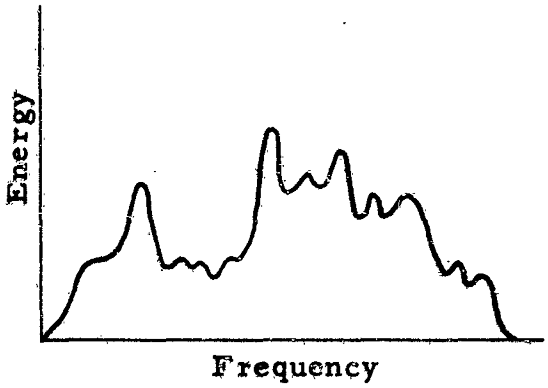

Fig. 1.1-B Typical Energy Distribution of Radio Interference

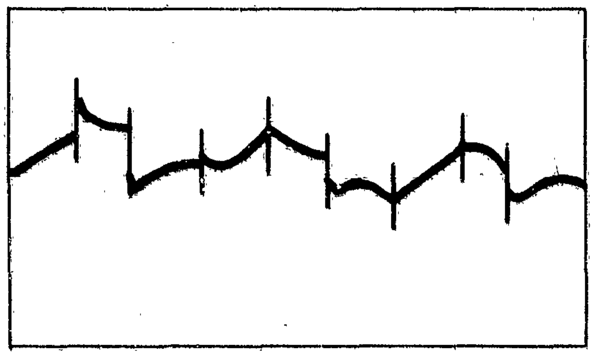

Fig. 1.1-C Oscilloscopic Trace of Commutator Interference

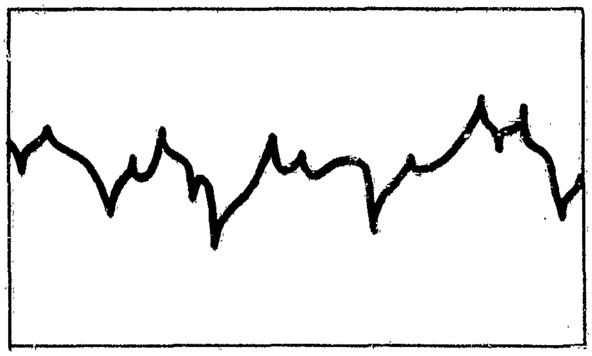

Fig. 1.1-D Oscilloscopic Trace of Typical "Random Noise"

The differences in wave form between interfering signal and "random noise" are clearly brought out in Figures 1.1-C and 1.1-D. Figure 1.1-C shows an oscilloscopic trace of a typical interfering voltage generated at the brushes of a motor. The periodicity, as well as the random fluctuations, are clearly evident. Figure 1.1-D is the trace of a truly random disturbance as obtained from a noise-generating diode exhibiting complete irregularity.

Despite these differences, certain results of the statistical analysis of "random

noise" find applications in the design of equipment for interference-free operation and will be pointed out and discussed at the appropriate places. A complete discussion of the methods by which these results were obtained is clearly beyond the scope of this book.

1.2 THE NATURE OF INTERFERENCE

In considering the final effect of the interference, i.e., its actual nuisance value, the most important factors are its magnitude in relation to that of the wanted signal and its position in the frequency spectrum. As far as the magnitude is concerned, it may be said that the nuisance value of an interfering signal varies directly as its magnitude, provided that it exceeds a certain minimum threshold value. This will remain approximately true even if the magnitude is measured not at the output but at the source or anywhere in the path of transmission of the interference from the source to the receiver. This is true because the transmission system is usually sufficiently linear to make the signal strength at any one point approximately directly proportional to the strength of the same signal anywhere else in the system.

As far as the position of the interfering signal in the frequency spectrum is concerned, no such simple statement can be made. For the final effect, the interference must either contain frequencies within the normal output range of the receiver, or it must be capable of making one or more stages of the receiver inoperative. In the first case, the final effect would be audio noise for an ordinary radio receiver, or it might be visual hash on a radar scope. In a navigational instrument, the final effect might be a false indication of an indicating needle. In the second case, the final effect would be the complete lack of an audio or visual indication. In either case, in order to produce this final effect at the output of the receiver, the interference at the input must contain frequencies within the band to which the receiver is sensitive. It is important to remember that <u>the band of frequencies to which the receiver is sensitive is much wider for interference than what is normally considered its "bandwidth"</u>. The attenuation of frequencies outside the normal transmission band is never infinite, and very often there is insufficient rejection of large interfering signals, even though their frequencies may be considerably removed from those the receiver is designed to accept. This is one way in which an interfering signal may affect a receiver, even though it has no frequencies within its nominal acceptance band.

There is another way in which an interfering signal, outside of the acceptance band of a receiver, may gain entrance. During transmission, <u>electrical signals may undergo one or several frequency translations</u>, i.e., they may combine with other signals in non-linear elements to produce entirely new frequencies. For example, the fundamental frequency output of a medium frequency transmitter may not itself fall within a band that is accepted by a low frequency receiver. But one of its higher harmonics may "beat" with the output of another transmitter in a non-linear element in such a way as to produce an interfering signal which does affect the receiver. Therefore, even though only a fairly narrow band of frequencies is effective at the input of the receiver because of its selectivity, frequencies in that band may be produced by entirely different frequencies at the source, and <u>it becomes necessary to treat signals of all frequencies as having potential nuisance value</u>.

<u>An interfering signal is always associated with a time-varying electric or</u>

magnetic field. If the variation of the field is sinusoidal, the resultant signal is completely specified by three quantities: its amplitude, frequency, and phase. More often, the variation will not be sinusoidal. In this case, an infinite number of parameters will, in general, be required for a complete description. This is evident from the following considerations: The time variation of the field intensity may be expanded either as a Fourier series or as a Fourier integral, the series being used when the variation is periodic and the integral when it is not. In the first case, the Fourier series will, in general, contain an infinite number of terms whose magnitudes and phases must be specific for a complete description of the series. In the second case, a complete description of the Fourier integral requires the specification of the amplitude and phase functions at all frequencies. Even though an infinite number of parameters is required for a complete description of the signal, usually only a finite number of terms or a limited range of frequencies need be considered because in all cases of practical importance the amplitudes of the high frequency components become too small to affect the receiver.

The physical interpretation of the above paragraph is that an arbitrary signal may be considered to consist of an infinite number of sinusoidal signals, superimposed. These sinusoidal signals are either of finite amplitude and occupy discrete frequencies, or they have infinitesimal amplitudes and a continuous frequency distribution. Such an analysis, usually called a "Fourier analysis", has the great advantage that if one deals with a linear network, for which the principle of superposition applies, the response of the network to any arbitrary input can be determined on the basis of a knowledge of the response of the network to sinusoidal inputs at all frequencies. This is the main reason why the consideration of non-linear networks leads to much greater difficulties. The response of a non-linear network cannot be found from an analysis of its behavior under excitation from sinusoidal sources.

The interfering signal, considered as a varying field, current, or voltage, is determined not only by the process of its generation, but also by the impedance into which the signal generator sends the impulse. Since the form of the resulting response (which is the only quantity that can be observed) is usually of greater interest than the hypothetical form of the generator output in the absence of any impedance, it is important to investigate the effect of various kinds of impedance on the form of any given impulse. To carry through such an investigation for all possible, or even all practically important, forms of impulses, is a prohibitive undertaking. A considerable insight into the problem may be gained, however, by concentrating on just one representative wave form. The particular form chosen is that of a rectangular pulse of unit area and of short but finite duration, because this form is often fairly closely approximated by actual interference pulses. Furthermore, a signal of any shape can be closely approximated by a succession of such pulses.

Considerable work is saved by utilization of the principle of duality. This principle states that, for certain network pairs, identical relations exist between the voltages, currents, impedances, and admittances, provided that all voltages, currents, impedances, and admittances in one network are replaced by the currents, voltages, admittances, and impedances, respectively, in the other. Network pairs which obey such laws are called duals of each other. Any planar network (i.e., one which can be projected onto a plane without any connections crossing one another) has a dual. The dual may be obtained by putting all series elements of one, in parallel in the other, and all parallel elements of one, in series in the other, by leaving

all resistances unchanged, by replacing all inductances with capacitances, and by replacing all capacitances with inductances. A similar principle of duality may be stated for fields instead of networks.

A distinction is sometimes made between network pairs having dual configuration and actual duals. In the first case, it is necessary only that the above-mentioned relations exist between types of elements without regard to their magnitude. For actual duals, there must also be a relationship between the magnitudes of corresponding elements, as shown in Figure 1.2-A, which gives an example of two networks that are actual duals of each other. Here k is an arbitrary real constant.

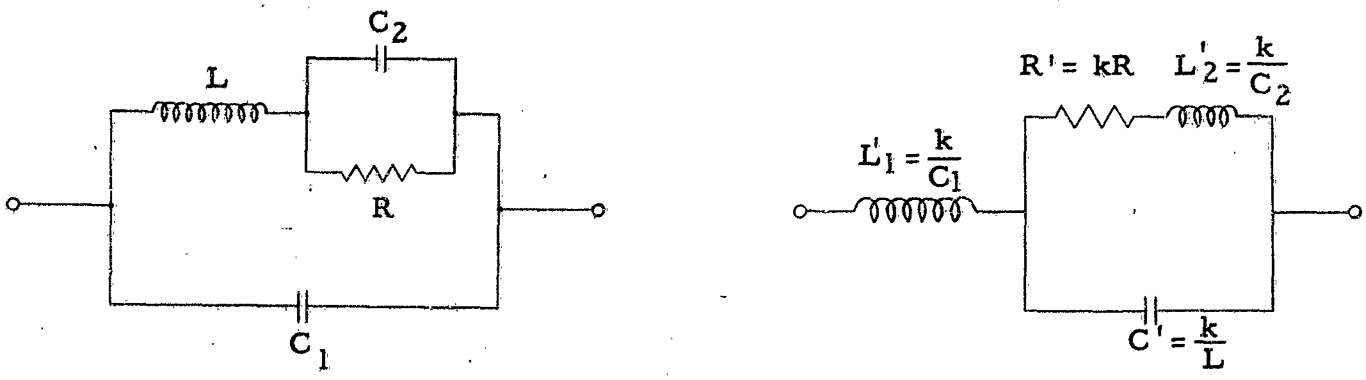

Fig. 1.2-A Dual Networks

By utilizing the principle of duality, it is found that the response of one network to a voltage impulse is the same as the response of its dual to the corresponding current impulse. For example, the current that results from applying a specified voltage to a series combination of resistance and capacitance has exactly the same form as the voltage that appears across a parallel combination of resistance and inductance when a current of the same specified time variation is sent into it. In this way, each separate analysis immediately yields two significant results.

Figure 1.2-B shows the responses of the simplest combinations of one, two, or three circuit elements to a short rectangular pulse. In each case the response shown is the current that will flow if a rectangular voltage pulse is applied across an impedance, Z, or the voltage across an admittance, Y, when a rectangular current pulse is flowing into Y. If the network Z consists of several branches in parallel, or the network Y has several elements in series, the result is simply a combination of the results shown, since the current in each branch or the voltage across each element may be found from Figure 1.2-B.

The actual interfering signal usually consists of a series of periodic or nonperiodic pulses similar to those shown in Figure 1.2-B. The responses will remain substantially as shown whenever the individual pulses occur so far apart that the energy of each pulse is practically dissipated before the beginning of the next one. (The cases without resistance, and hence without dissipation, need not be considered here since they cannot be realized in practice.) If, however, the pulses follow each other so closely that the initial conditions of each pulse are affected by the previous one, then the transient caused by one pulse is superimposed on the transient of the previous pulse, which did not have time to die out, and the resulting wave form may have very little resemblance to the response shown for a single pulse. In the case

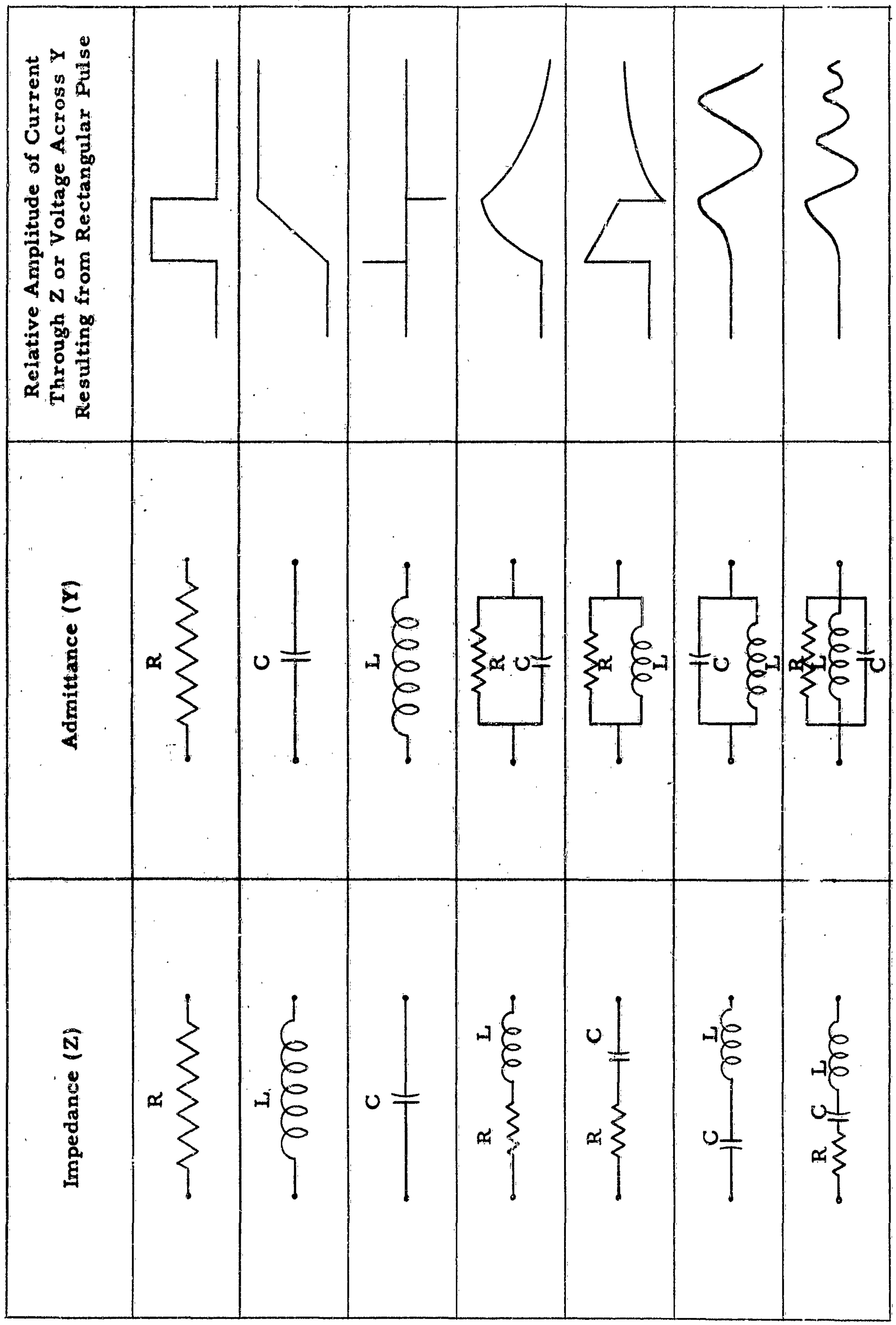

Fig. 1.2-B Response of Single Networks to Rectangular Pulse

of a resonant circuit, for example, the output wave will deviate more and more from the damped sine wave shown in the last diagram of Figure 1.2-B as the ratio of the resonant frequency to the pulse repetition rate becomes smaller.

The results of Figure 1.2-B may be obtained in two ways: (a) determine what happens in the various elements as voltage (or current) is suddenly applied to them, or (b) consider the response of the network to sinusoidal excitations. In the first method, attention is focused on the way in which a capacitance stores charges, an inductance introduces inertia effects, and a resistance causes dissipation of energy. In the second method, a Fourier analysis of the rectangular pulse must be performed. This latter point of view shows that, as the frequency of the applied sinusoidal voltage increases, the current in a capacitive circuit will increase, the current in an inductive circuit will decrease, and the current in a resonant circuit will increase at first, reach a maximum at the resonant frequency, and then decrease. If the frequency spectrum of the exciting pulse is known, the frequency spectrum of the response may then be predicted, and the response itself may be determined.

For simple circuits, the first method is not only sufficient, but actually more enlightening, because it is easier to understand what actually goes on in the network. But for more complicated networks and general analyses, the second approach is indispensable. An analysis of this type proceeds as follows:

An arbitrary signal, which may bear no resemblance to the rectangular pulse used as an example before, is given as a function of time, $f = f(t)$. Also, there is given a network whose response to sinusoidal excitation is specified at all frequencies, in the form of a complex function $G = G(\omega)$ where ω is the angular frequency in radians per second. The task at hand is to determine the response of the network. The procedure is to find the Fourier transform of $f(t)$, designated as $F(\omega)$:

$$F(\omega) = \frac{1}{2\pi} \int_{-\infty}^{+\infty} f(t) \; e^{-j\omega t} \, dt \tag{1}$$

and to obtain the Fourier transform of the response simply as the product of GF. If the response is indicated by $h = h(t)$ and its Fourier transform by $H = H(\omega)$, then

$$H(\omega) = G(\omega) \, F(\omega) \tag{2}$$

and

$$h(t) = \int_{-\infty}^{+\infty} H(\omega) \; e^{j\omega t} \, d\omega = \int_{-\infty}^{+\infty} G(\omega) \, F(\omega) \; e^{j\omega t} \, d\omega \tag{3}$$

The square of the absolute value of $F(\omega)$, $|F(\omega)|^2$, is proportional to that portion of the energy of the signal $f(t)$ that is associated with the angular frequency ω. Therefore, $|F(\omega)|^2$ will simply be called the energy distribution of $f(t)$.

Listed below are certain relationships between the properties of the function

*The form $F'(\omega) = (1/2\pi) \int_{-\infty}^{+\infty} f(t) \, e^{j\omega t} \, dt$ can also be used as a Fourier transform. It will be recognized that F' is simply the complex conjugate of F. The form used in the text is preferred in most of the literature.

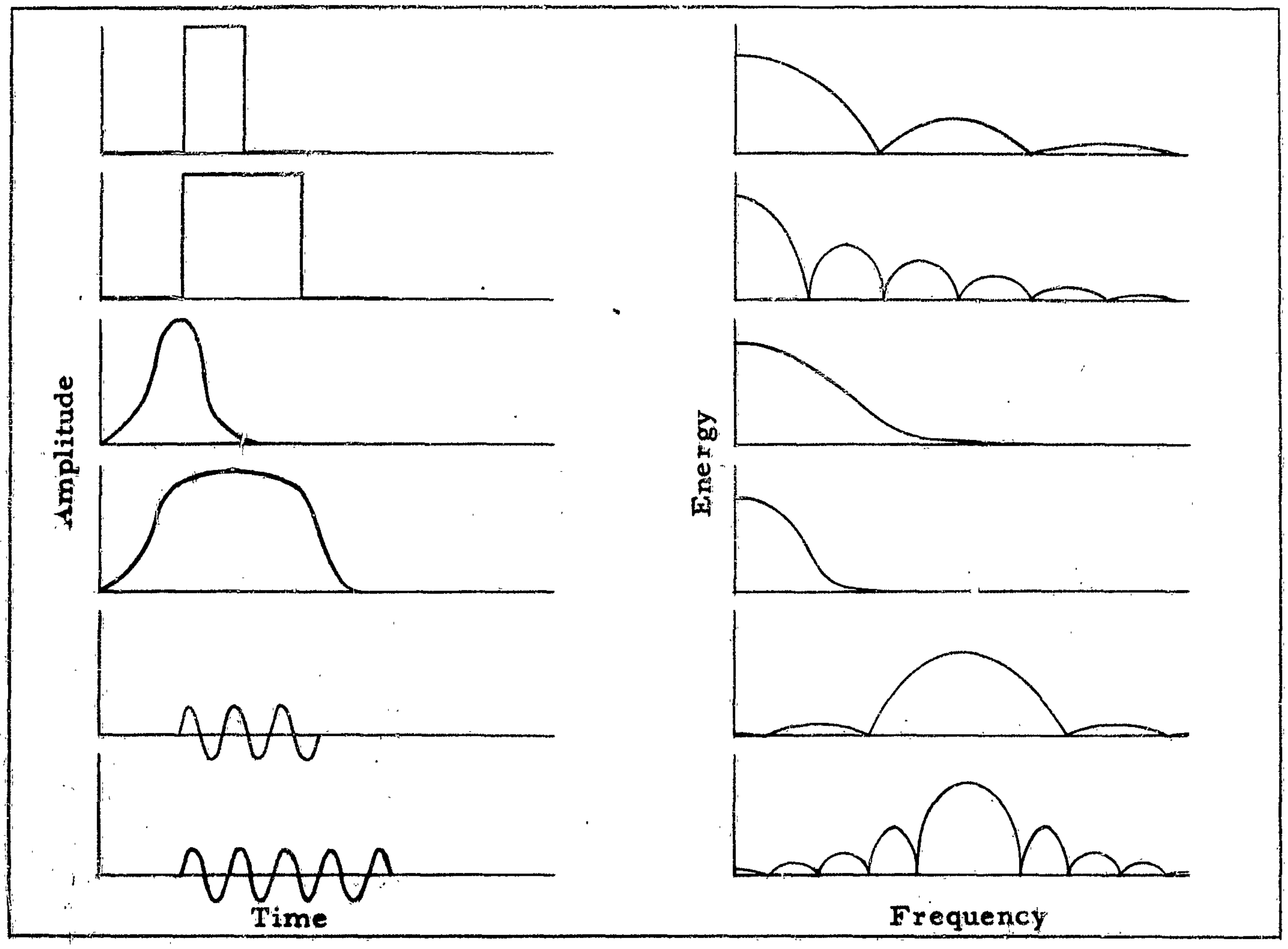

Fig. 1.2-C Energy Distributions Showing "Reciprocal Spreading"

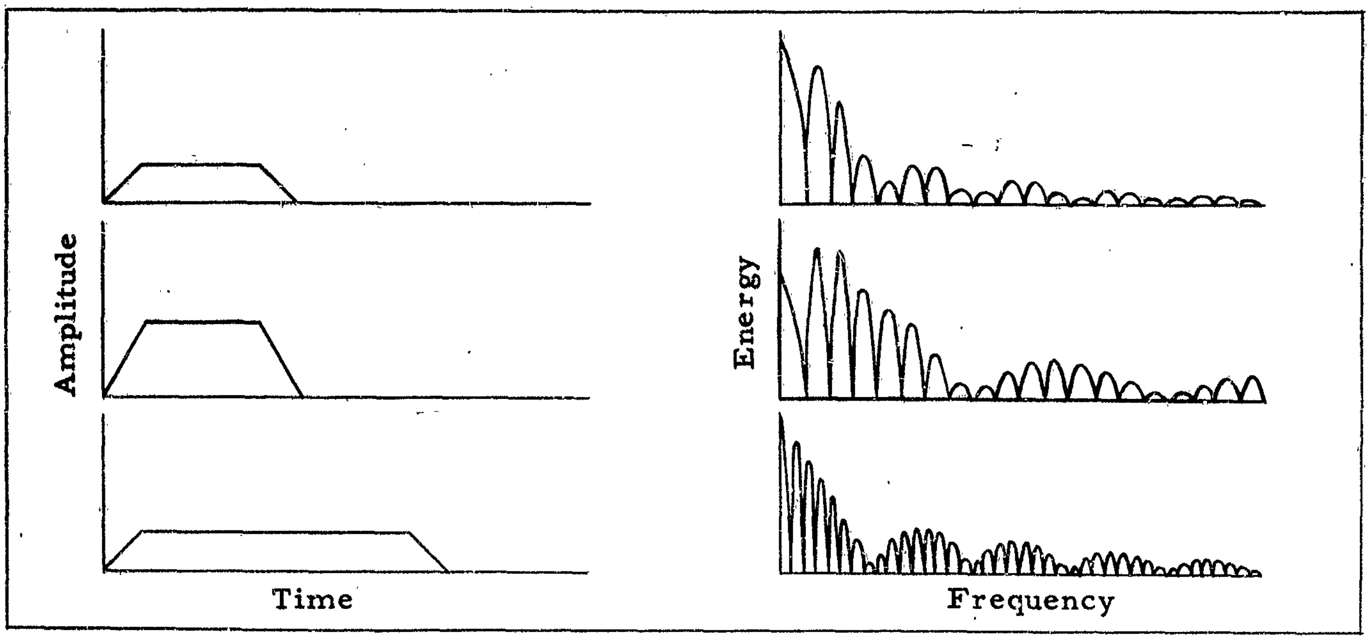

Fig. 1.2-D Energy Distributions Showing Effects of Steepness of Rise and Duration of Pulse

f(t) and its frequency distribution $F(\omega)$, which are important for the analysis of interference.

(a) Reciprocal Spreading. For a pulse of finite duration, the spread in frequency is roughly inversely proportional to its duration. Examples of this are shown in Figure 1.2-C.

(b) Effect of Steep Wave Fronts. For any pulse, the sharper its rise or fall, the greater is that portion of its energy which is concentrated in the high frequency components. This effect leads to the conclusion that only those pulses which have sharp wave fronts will cause appreciable interference at high frequencies. An example of this is shown in Figure 1.2-D. The computations for these, and the following examples, are given in detail in Appendix IV.

(c) Effect of Duration of Pulse. For any pulse, the longer its duration, the greater is that portion of its energy which is concentrated in its low frequency components. An example of this is shown in Figure 1.2-D.

These rules are qualitative rather than quantitative, since it may be impossible to assign exact values to the "duration" of a pulse which decays to zero exponentially, or to the "spread" of a function that goes to zero like a damped sine wave.

The method of Fourier analysis outlined above is extremely powerful and finds many practical applications. In this book it will be used (1) to help formulate the impedance approach in Appendix V, (2) to predict the response of receivers to the most common types of interference, and (3) to analyze the changes which the interfering signals may be expected to undergo during their transmission from the source to the receiver (see Paragraph 1.4). For example, the Fourier analysis of the interference generated by a particular motor, may reveal a strong peak in the vicinity of, say, 2 megacycles per second, as shown in Figure 1.2-E. It will then be necessary to design a filter whose attenuation characteristic has somewhat the same

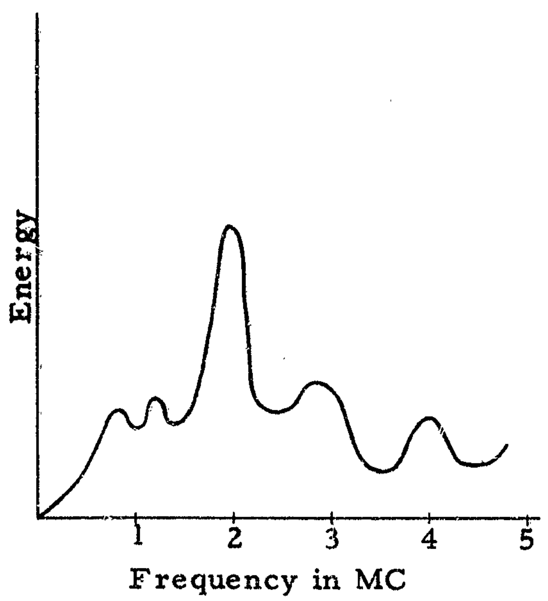

Fig. 1.2-E Typical Frequency Spectrum of Motor Interference

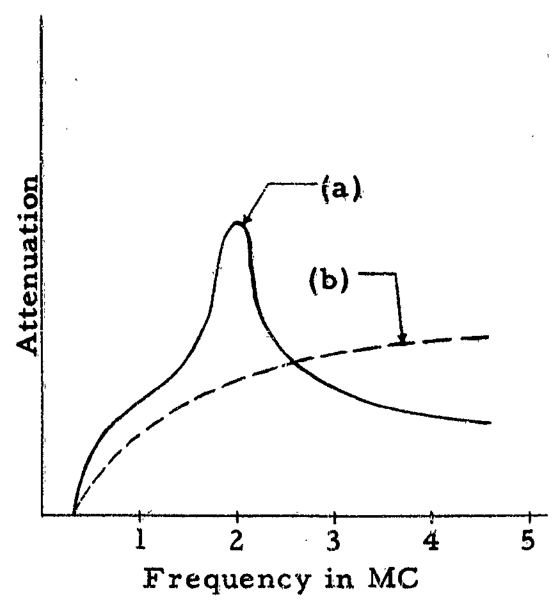

Fig. 1.2-F Attenuation of Filter for use with Motor of Fig. 1.2-E

shape as is shown in Figure 1.2-F. This could be accomplished by use of a series-derived M-type filter giving an attenuation as shown in Figure 1.2-F, curve (a). A low pass filter, of the constant-K type, would probably not be sufficient to suppress the interference in the neighborhood of 2 megacycles, as shown in Figure 1.2-F, curve (b). For further information in design values for filters, see Appendix VII.

1.3 THE ORIGIN OF INTERFERING SIGNALS

It was said before that interference is always associated with varying electric or magnetic fields. This is true, but the varying fields, themselves, are not the original source of the interference. What, then, causes varying electric or magnetic fields?

An electrostatic field exists wherever there are charges. If charges are stationary, the field remains constant and there can be no disturbance. Whenever the charges are moving, there is a magnetic field, but, as long as the motion is uniform, the magnetic field is constant and again there can be no disturbance. It follows that before there can be any interference, there must be a non-uniform motion of charges, i.e., a varying current. And, indeed, as the various causes of interference in electrical equipment are investigated, one finds that in each case, a varying current is responsible. Once the current is considered, the charge may be forgotten, because a moving charge is a current, and one may base the discussion on one or the other, but not on both. It may be said, then, that the condition for no interference is

$$\frac{di}{dt} = 0 \tag{4}$$

where i stands for current and d/dt for differentiation with respect to time. This expression is extremely simple, yet important to keep in mind as the ideal for which to strive. In some cases, a varying current is essential to the operation of the equipment, as, for example, in an alternator. In this case, Equation (4) cannot be satisfied. But even then Equation (4) remains valid, provided one lets i refer to that portion of the total current which remains after subtracting the desired current. Then the equation is not the condition for "no interference", but the condition for "no more interference than is absolutely necessary for the proper functioning of the equipment".

To determine the actual origin of the interference, the causes of a varying current must be located. The object of this procedure is not to make the obviously absurd attempt of finding the "origin of the origin of the origin", but rather to seek the points at which the application of corrective measures is most practical and effective. It is found that, in some cases, these points are reached when mechanical rather than electrical considerations are involved. In other cases, these points are reached when electronic means of current generation are involved.

What causes a varying current? A current may flow either in a conductor (conduction current), in a gaseous dielectric through which charged bodies or particles are moving (convection current), or in a dielectric void of free charges (displacement current). Displacement currents are negligible at all frequencies at which the considerations of ordinary circuit analysis are valid and need to be considered

only in connection with radiation and other phenomena associated with high frequencies. Convection currents occur in electron tubes and in electric arcs and sparks. Also, any charged body in motion constitutes a convection current. Since an insulator will not carry an appreciable amount of net charge, the only bodies that need to be considered in this connection are conductors While it is true that any motion of a charged conductor of finite dimensions will result in a varying field and, therefore, is a potential source of interference, it may be stated that such interference is very rare and not likely to be encountered in aircraft. The currents which are important as sources of interference are the first mentioned convection currents, i.e., those consisting of moving electrons or ions, and the conduction currents.

These important currents may always be computed by an application of the basic equation $i = E/Z$, in which the current i is considered the effect produced by the electromotive force E against the opposition of the impedance Z. From this it may be concluded that variations of current may be caused either by a varying electromotive force or by a varying impedance. The final answer, then, to the question asked in this Paragraph 1.3, is that there are two basic processes in which interference may originate; one is the generation of varying electromotive forces, the other the variation of impedances.

1.3.1 THE GENERATION OF VARYING ELECTROMOTIVE FORCES

There are three processes of generation of electromotive forces which will be discussed in the following paragraphs; generation by mechanical means in rotating machinery, generation by switching action in vacuum tube oscillators, and generation of equivalent voltages in non-linear impedances. These three are, by far, the most important sources of interference in aircraft under the heading of this paragraph. The generation of electromotive forces by chemical means in batteries may also be a source of interference, but such interference is extremely minute and not important for this book. The generation of electromotive forces by friction and other causes on the outside of aircraft is important, but will be treated separately in Section 4. Voltages induced by the action of mutual inductance need not be considered, since the voltage generating the inducing current is a more elementary consideration.

1.3.1.1 ROTATING MACHINERY

In all rotating machinery, there is a relative motion between a set of conductors and a magnetic field. An electromotive force, E, is induced in the conductors, which may be computed according to the basic law,

$$E = B L v \tag{5}$$

in which B is the magnetic flux density, L the effective length of the conductor perpendicular to the field, and v the component of relative velocity perpendicular to L and B. The quantity B is varying in all cases encountered in aircraft, because it is not practical to have the motion take place in a constant field. Ideally, in an alternating current machine, the variation is such that the generated voltage is a pure sine wave. Ideally, in a direct current machine, the variation is such that the generated voltage, as measured at the terminals, is constant while the brushes slide on any one commutator segment, as well as when they slide to the next segment. Actually, deviations from the ideal are always present in both types of machines. The

peripheral velocity, v, of the conductors will not be exactly constant, and the effective length, L, of the conductors will not always be exactly the same for all conductors. Thus, irregular variations occur in all factors on the right of the above equation, and, as a result, the generated electromotive force, E, will always contain undesired variations. It is the task of the machine designer to produce a machine in which these variations are as small as possible.

In a direct current machine, an additional difficulty arises. Even if the ideal is approached very closely and the output voltage appearing at the brushes is free from any "ripples", the voltages induced in the conductors must jump abruptly from one constant value to another each time the brushes pass from one commutator segment to the next, so that the current in the armature will show very rapid variations and therefore be very rich in harmonics, even though the external current may be free of them. This is one of the reasons why direct current machines are more troublesome from the point of view of radio-interference than alternating current machines. (The other is that the process of commutation itself practically always introduces harmonics into the external current, as will be pointed out in Paragraph 1.3.2.2.) Obviously, the design engineer must develop rotating electrical machinery with the best possible mechanical and electrical characteristics in order to minimize the generation of interference.

1.3.1.2 VACUUM TUBE OSCILLATORS

An alternating electromotive force may also be generated in a vacuum tube oscillator. The frequencies required for the normal operation of many types of electronic equipment, e.g., the frequencies generated in the local oscillators of superheterodyne receivers, often cause interference in other equipment even when the oscillator is an ideal one and generates a perfect sine wave. The interference caused by the fact that the oscillator is not an ideal one, will be discussed later. Here, attention is drawn to the fact that <u>the oscillators are likely to be sources of interference by virtue of the very function they are called upon to perform.</u>

1.3.1.3 NON-LINEAR IMPEDANCES

Impedances which vary in time cause varying currents, as will be discussed in the next paragraph. But <u>impedances that vary with the current through them or with the voltage applied across them (so-called "non-linear" impedances), act like equivalent generators of varying electromotive forces.</u> This may be shown as follows. Cons .r the circuit of Figure 1.3.1.3-A. The generated voltage of the generator is $e_0 = E_0 \sin \omega t$, and the non-linear impedance Z is assumed to be a function of the current i, $Z = f(i)$, which can be developed into a Taylor series:

$$Z = Z_0 + A_1 i + A_2 i^2 + \ldots \tag{6}$$

where the coefficients Z_0, A_1, A_2, etc. are assumed to be known. Then the current is

$$i = \frac{e_0}{Z} = \frac{E_0 \sin \omega t}{Z_0 + A_1 i + A_2 i^2 + \ldots} \tag{7}$$

This implicit equation for i may be solved explicitly by the method of successive approximations if the assumption is made that A_1 and A_2 are small and all higher terms are negligible. Denoting successive approximations to i by i_0, i_1, etc., one has:

$$i_0 = \frac{E_0}{Z_0} \sin \omega t \tag{8}$$

$$i_1 = \frac{E_0 \sin \omega t}{Z_0 + A_1 \frac{E_0}{Z_0} \sin \omega t + A_2 \left(\frac{E_0}{Z_0}\right)^2 \sin^2 \omega t} \tag{9}$$

$$\simeq \frac{E_0}{Z_0} \sin \omega t \left(1 - A_1 \frac{E_0 \sin \omega t}{Z_0^2} - A_2 \frac{E_0^2 \sin^2 \omega t}{Z_0^3}\right) \tag{10}$$

It is not necessary to go any further in order to see that the second term in Equation (6) gives rise to a term that varies as $\sin^2 \omega t = (1/2)(1 - \cos 2\omega t)$, and that the third term in Equation (6) gives rise to a term that varies as $\sin^3 \omega t = (1/4)(3 \sin \omega t - \sin 3\omega t)$. Therefore, the current will contain a second and third harmonic that were not present in the original source. It is possible, in general, to replace any non-linear impedance by a linear impedance in series with one or more generators whose frequencies are harmonics of the actual source frequency. Thus, the circuit of Figure 1.3.1.3-A, which contains a non-linear impedance, is equivalent to the circuit shown in Figure 1.3.1.3-B, which contains only linear impedances, but has additional signal generators.

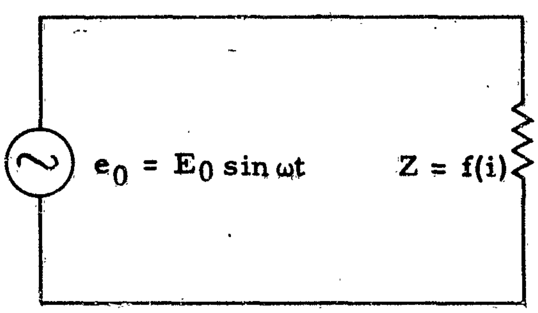

Fig. 1.3.1.3-A Circuit Containing a Non-linear Impedance

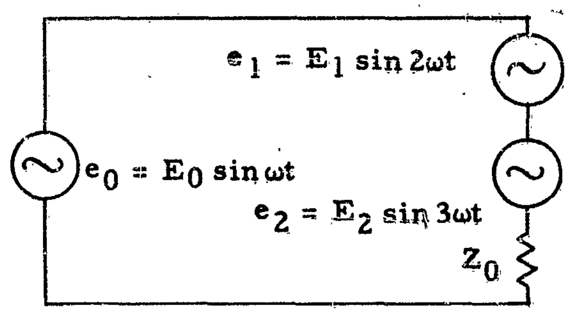

Fig. 1.3.1.3-B Circuit with Linear Impedance Equivalent to the Circuit of Figure 1.3.1.3-A

Still greater complications arise in cases where the impedance Z is a function of both the current i and its rate of change di/dt. This happens, for example, in a non-linear inductance. While the mathematics involved are more complicated, the net results are essentially the same. It is still possible to replace the non-linear impedance by a linear one in series with a number of harmonic generators.

Thus, it is seen that any non-linear impedance is a possible source of interference.

1.3.2 VARIATION OF IMPEDANCE

The variation of impedance to the flow of current is one of the most important causes of undesired current variations. Usually the variation is one of the resistance, but variations of inductance and capacitance may be troublesome also. The most extreme case of variable impedance is a switch, where the impedance changes from some finite value to infinity or vice versa. Less extreme cases occur, where the impedance changes, more or less suddenly, from one finite value to another; but from the discussion in Paragraph 1.2 it is clear that very slow and gradual changes will not result in appreciable interference. On the other hand, very sudden changes, even though small, may cause a large number of undesired harmonics of appreciable amplitude.

The most important cases of trouble-causing impedance variations are discussed in the following paragraphs.

1.3.2.1 BRUSHES

Electrical contact between circuit components in relative motion is usually made by means of brushes that ride on slip rings or commutators. Due to the mechanical action of friction, both surfaces wear down gradually, although the brushes wear down faster than the harder metallic surface of the slip ring. The process of brush wear, which is gradual on a macroscopic scale, is actually very irregular and of random nature on a microscopic scale. Fairly large carbon particles are torn loose and either ejected or burned, so that the contact resistance, which depends both on the pressure and on the actual contact area, is subject to sudden random variations of considerable magnitude. In addition to this, vibrations may be set up in the mechanical system of the brushes, springs, and brush holders, which cause the pressure, and thus the resistance, to vary. In extreme cases, the variations of pressure may be so great that the brush bounces completely off the metal, thus causing a true switching action. These causes combine to make all brushes a serious source of interference.

1.3.2.2 MECHANICAL SWITCHES AND COMMUTATORS

When a switch is opened or closed in an electrical circuit, the impedance in the branch containing the switch changes suddenly from practically zero to infinity, or vice versa. The currents and voltages in the circuit must readjust themselves, and whenever the circuit contains inductances or capacitances, this adjustment cannot take place instantaneously, because the energy stored in the magnetic field of the inductance or electric field of the capacitance cannot change instantaneously. There is a brief time interval during which the readjustment takes place, and the changing currents and voltages during this interval are called transients. As was pointed out in Paragraph 1.2, any transient may cause radio interference, and therefore any switch is a potential source of interference.

In addition to transients produced during the normal functioning of a switch, severe interference may be created due to arcing between the contactors of the switch, immediately before making or after breaking physical contact. If the voltage between contactors is high and the process of closing the switch comparatively slow, there may be a time when the points of contact are separated by a distance shorter

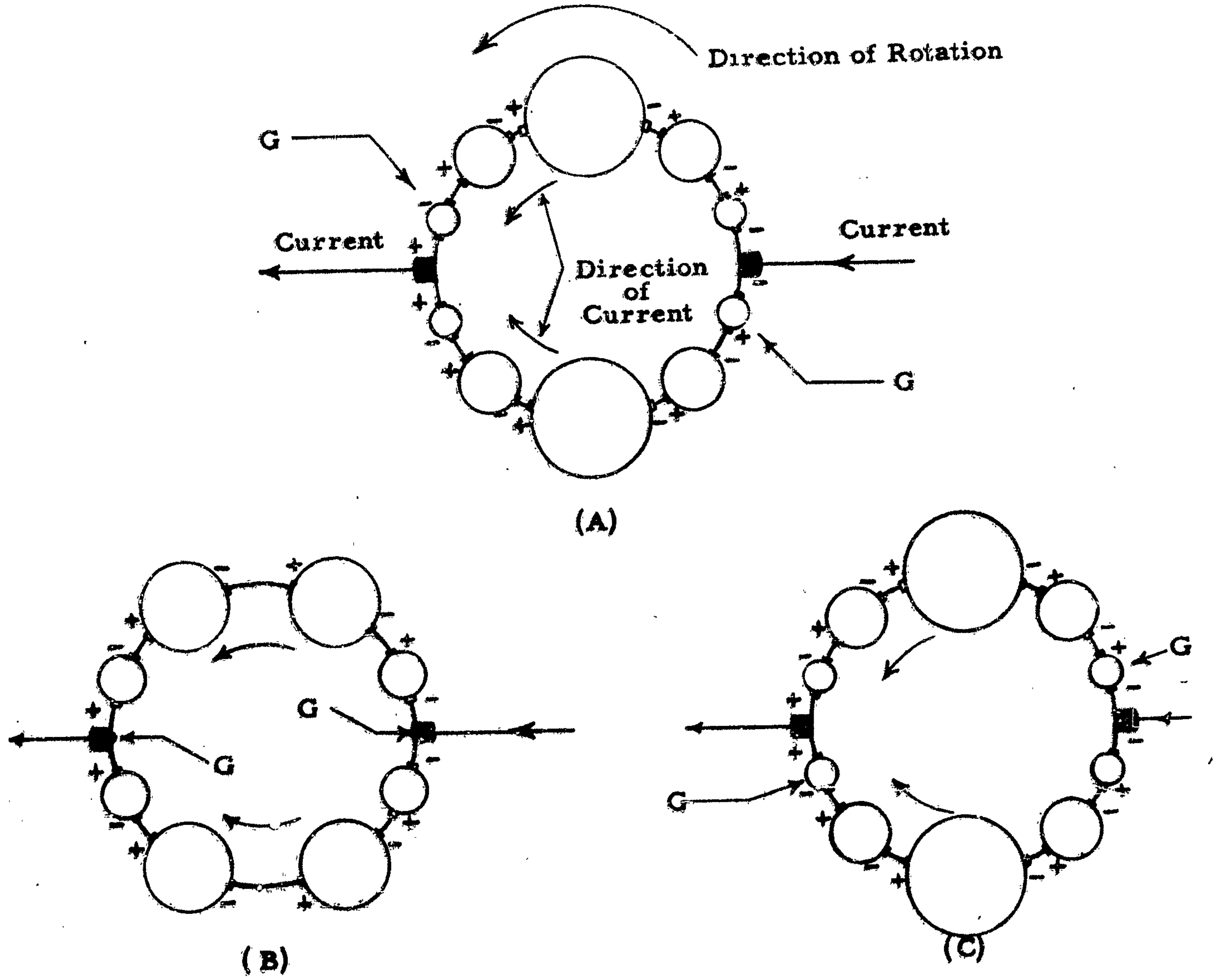

Fig. 1.3.2.2-A, B, C Illustration of Switching Action of a Commutator.
(Individual Armature Conductors Represented by Generators)
(Magnitude of Induced Voltage Indicated by Size of Circles)

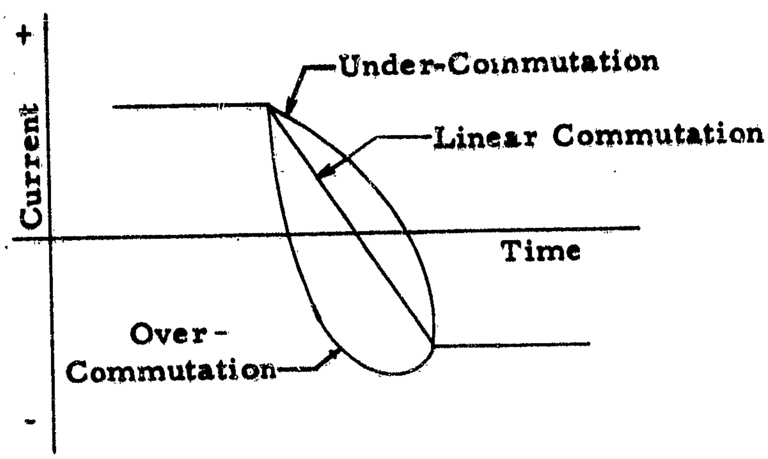

Fig. 1.3.2.2-D Variation of Current in Coil Undergoing Commutation

than that required for breakdown of the insulation between them; and if the time required for ionization is shorter than that required to close the switch fully, arc-over will occur. A somewhat similar, but more frequent, phenomenon is arc-over between contacts during the process of separating them, providing that the circuit has enough inductance to induce the voltage necessary to maintain the arc. In either case, the resistance of the arc varies rapidly through a wide range of values, and, hence, varying currents, rich in interfering frequency components, are produced. Since an arc tends to maintain a current that is decreasing to zero, arcing actually decreases the severity of the transients by decreasing the rate of change of the current. But little is gained because the arcing itself produces severe interference. The result of normal switching transients, whether or not accompanied by arcing, is the familiar clicking sound so often heard in the earphones of radio receivers during switching processes.

One of the most important examples of an interference-producing switching process, is commutation in rotating electrical machines. The function of a commutator is to switch the output terminals from one set of terminals on the armature to another in such a way as to keep the current and generated voltage as constant as possible. The switching action of a commutator is explained with the aid of Figures 1.3.2.2-A, B, and C. In each diagram, the individual armature conductors are represented by generators, the size of each generator indicating the magnitude of the voltage induced in it. In Figure 1.3.2.2-A, the generators/(coils) G are about to undergo commutation. The voltage induced in them is very small, but one half of the full line current is flowing through them in the direction indicated by the arrow. In Figure 1.3.2.2-B, the same generators (coils) have moved under the brushes and are now undergoing commutation. They are short-circuited and the voltage induced in them is ideally zero while the current is changing to a new value. In Figure 1.3.2.2-C, commutation in generators G has been completed. Now a small voltage in the opposite direction is induced in them, and the current in them is again one half of the full line current, but the current now flows opposite to its previous direction. During commutation, the current must change from some positive value to an equal negative value. The variation may be as shown in one of the curves of Figure 1.3.2.2-D, which are called over-commutation, under-commutation, and linear commutation, respectively.

This analysis makes clear that, even under ideal conditions, direct current machines with commutators must be a source of interference in three distinct ways: (1) the current in the coil undergoing commutation must change rather rapidly, (2) the voltage generated must vary as the voltage induced in each coil varies with its position in the magnetic field, and (3) the total armature impedance between brushes must change as some of its coils are short-circuited by the brushes. In addition, there are many opportunities for interference to be generated due to deviations from the ideal, such as the voltage induced in the coil undergoing commutation not being exactly zero, or arcing at the brushes.

1.3.2.3 ELECTRONIC DEVICES

In many applications, a vacuum or gas-filled electron tube is used to produce switching action. These tubes are particularly useful as generators of non-sinusoidal wave forms and find wide applications as pulse generators, modulators, and oscillators. In these applications, the generation of harmonics is desired and is essential

to the proper operation of the device. It is not surprising, then, that these devices also rank high as interference generators.

The basic circuit of an oscillator of this type (a so-called "relaxation oscillator"), is shown in Figure 1.3.2.3. When the battery switch is closed, the condenser charges through a resistance until the voltage across the tube exceeds the breakdown voltage. When that point is reached, the condenser discharges very rapidly through the low impedance of the tube until the voltage across the tube drops to a value below that necessary to maintain the arc. After that, the tube impedance becomes infinite, and the entire cycle repeats. All electronic pulse generators or non-sinusoidal oscillators use this basic principle, although, in practice, the circuits will be more complicated. In particular, much more flexible control may be achieved by using a triode, and applying a control voltage to the grid, instead of relying entirely on the plate voltage to cause breakdown ("closing" of the switch) at the desired instant. If a triode is used, vacuum tubes may be substituted for gas tubes because a sufficiently large negative control voltage on the grid will "cut off" the plate current, thus opening the switch.

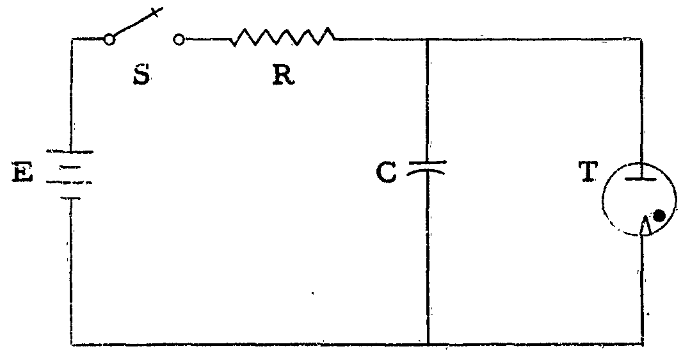

Fig. 1.3.2.3 Simple Relaxation Oscillator

An undesired variation of impedance also takes place in vacuum tubes used as sinusoidal generators. Due to the non-linearity of the tube characteristics, there will always be some harmonics generated in an electronic oscillator, together with the desired frequency. In fact, this effect is unavoidable, though it may be kept quite small, because an oscillator is essentially a non-linear device. Were it linear, the amplitude of oscillation would be unstable, all values being equally possible. It is the non-linearity of the tube that provides the stabilizing effect, by varying the impedance in such a manner as to counteract any small changes of amplitude which may occur. Thus, no vacuum tube oscillator produces a perfect sine wave, although in practice the harmonic generation can be kept extremely low. The larger the power output of the tube, the more difficult it becomes to keep the harmonics small. Therefore, it is usually the last stage of a transmitter, or similar electronic device, which is most likely to produce interference. However, the sensitivity of modern receivers is such that even very weak interfering signals may be very troublesome. Therefore, it becomes necessary to watch carefully all stages of transmitters and eliminate all harmonic generation, as much as possible.

1.3.2.4 ARCING

When the electric field intensity in a dielectric between two conductors exceeds the breakdown strength of the dielectric, an arc occurs, resulting in very rapid and

large variations of the impedance of the path between the conductors. The rapidity of impedance variation depends on the properties of the dielectric, as well as on the external circuit. In the case of a gas, it is the ionization and de-ionization time that is most important. Usually, the first is much shorter than the second; hence, the most rapid changes and the most severe interference take place during the actual breakdown of the gas. Arcing in solid or liquid dielectrics is not considered here, because its occurrence would be classified as a failure of the system and, therefore, would not be expected anywhere during normal operation. While not all gaseous discharges are arcs, it is not necessary to distinguish here between the various types of discharges. The word "arc" will be used for all types of electrical discharges discussed in this book.

Arcing occurs both by design and as an undesirable by-product. Its occurrence during switching processes has already been mentioned. It may also occur when a potential difference is developed between two structural members that are almost, but not quite, in contact. If the distance between contact points is variable, due to shock or normal vibration in the aircraft, there is the additional effect of the varying capacitance, which introduces a varying impedance, even in the absence of an arc. Since structural members in aircraft are universally used as ground return paths for power currents, and sometimes also for control and communication currents, it is clear that lack of good electrical contact at any point will be a serious source of interference.

Because arcing always produces large transients, special precautions must be taken when arcing occurs by design, as in the ignition system or the transmit-receive box of a radar set. These transients cause a large amount of radio interference unless they are prevented from reaching the receiver.

Arcing may also occur between external parts of the aircraft and the surrounding atmosphere. This phenomenon, called corona discharge, will be discussed in detail in Section 4.

1.4 CONSIDERATION OF THE ENTIRE AIRCRAFT AS A SINGLE NETWORK

In order for any interference to become effective, it must, in some way or other, reach the receiver. Within the airframe there is rarely a single mode of transmission from the point of generation to the receiver, i.e., transmission is rarely entirely by radiation, or entirely by inductive coupling. In subsequent paragraphs, such single modes will be described and analyzed, but in this paragraph, emphasis is placed on the fact that the entire aircraft, including all its equipment and wiring, must be considered as one entity in any complete analysis of interference. Frequently, one particular mode of transmission will be more important than the others, but there is always great danger that, in over-emphasizing this one mode, important interactions with other modes will be completely overlooked. Often these interactions modify the original mode to such an extent as to make the isolated analysis of that mode alone completely invalid.

For purposes of analysis, then, the entire aircraft must be considered as a four terminal transmission network, characterized by four parameters: (1) a single complex input impedance at the terminals to which the interference source is connected, (2) a single complex output impedance at the terminals to which the receiver

is connected, and, (3) and (4) two transfer impedances or transfer constants. These last two will be different, in general, but will be equal in the special case of greatest practical importance, when the network contains only bilateral elements.

As was already pointed out in Paragraph 1.2, the character of a signal is determined not only by the process of its generation, but also by the internal impedance of the source, as well as the impedance into which the source is operating. This statement emphasizes that source, transmitting network or medium, and receiver cannot be treated independently, but must be considered as one unit. This idea is referred to, for brevity, as the "impedance concept", because, as far as the source is concerned, the entire aircraft and receiver can be replaced by a single complex impedance, and, as far as the receiver is concerned, the entire aircraft and source can be replaced by a generator in series with a single complex impedance.

The consideration of the aircraft as a four-terminal network is adequate for most applications, but lacks generality in two respects. As will be discussed later (see Paragraph 1.6), the concepts of networks and terminals become meaningless at frequencies for which the wave length becomes comparable in magnitude with the dimensions of the equipment and wiring involved. The term "network" may be generalized so as to acquire meaning for this case also, but it will be more convenient to talk about "transmitting media" which are coupled to the source and the receiver and to each other through electric or magnetic fields, or through mutual impedances that cannot be localized at any terminals.

The other lack of generality lies in that this impedance concept does not apply readily in the presence of non-linear impedances. If the principle of superposition does not apply, consideration of a four-terminal network is not sufficient, because of the possible interactions between the interfering signal under consideration and some other signal, which may be a desired or another interfering signal. Fortunately, non-linear elements are vary rare in the transmitting network, and occur mostly within the receiver itself. They will be discussed in that connection in Paragraph 1.7. For the present, it is assumed that only linear elements need be considered, and, therefore, the consideration of four terminals is adequate.

The general theory of four-terminal transmission networks is extensively treated in several textbooks. There, it is shown that any linear, passive, four-terminal network is completely specified by four complex parameters, which, in general, are functions of frequency. Several sets of four may be used, but the two image impedances and the two transfer constants are the most convenient ones for the present purposes. The image impedances are those impedances which must terminate the network in order to have an impedance match (i.e., equal impedances looking into and out of the network; not the conjugate match required for maximum power transfer) at each pair of terminals. The transfer constants are one half of the logarithm of the ratios of volt-amperes in to volt-amperes out when the network is terminated in its image impedances. For a network containing only bilateral elements, three, instead of four, independent parameters suffice to specify the network, since the two transfer constants become equal. If, in addition, the network is symmetrical, the image impedances are equal also, and only two independent parameters are left.

Consider, now, the circuit of Figure 1.4. The transmission network is linear and bilateral, but not necessarily symmetrical. It has image impedances, Z_{I_1} and

Z_{I_2} and a transfer constant θ. A generator of voltage E and internal impedance Z_S, is connected to the input terminals, and a load Z_R is connected to the output terminals. It is assumed that the impedances are not matched on either side, so that $Z_S \neq Z_{I_1}$ and $Z_R \neq Z_{I_2}$. Then the currents in the generator and in the load impedance are given by:

$$I_1 = \frac{E}{Z_S + Z_{I_1}\left[\dfrac{1 + F_r e^{-2\theta}}{1 - F_r e^{-2\theta}}\right]} \tag{11}$$

where $F_r = (Z_R - Z_{I_2}) / (Z_R + Z_{I_2})$, and

$$I_2 = \frac{E}{2Z_S} \cdot \frac{2Z_S}{Z_{I_1} + Z_S} \cdot \frac{2Z_{I_2}}{Z_R + Z_{I_2}} \cdot e^{-\theta} \cdot \sqrt{\frac{Z_{I_1}}{Z_{I_2}}} \cdot \frac{1}{1 - \dfrac{(Z_{I_1} - Z_S)(Z_{I_2} - Z_R)}{(Z_{I_1} + Z_S)(Z_{I_2} + Z_R)} e^{-2\theta}} \tag{12}$$

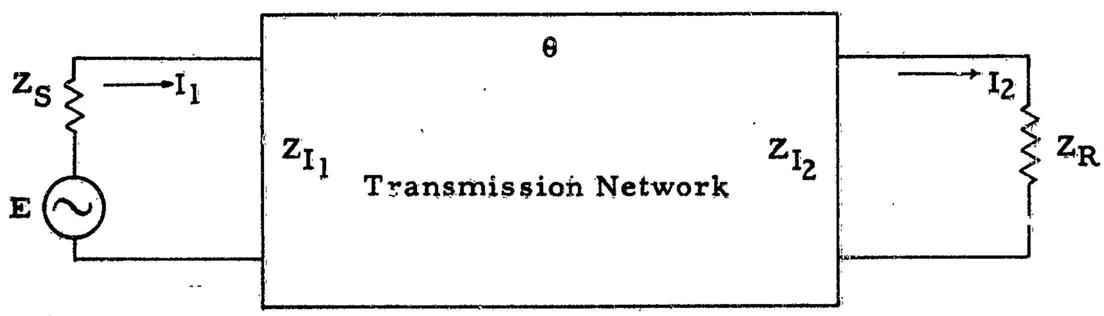

Fig. 1.4 Generalized Four Terminal Transmission Network

The expression for I_1 in Equation (11) means that the input current may be obtained by taking the ratio of the voltage E to the total impedance of the circuit. Hence, the second term in the denominator must be the impedance of the network when terminated in Z_R. The various factors in the expression for I_2 have very simple interpretations also. The first, $E/2Z_S$, is the current that would flow if the source were simply connected to a matching impedance Z_S. The second term, $2Z_S/(Z_{I_1} + Z_S)$, called the transmission factor, shows that this ideal current is modified in going from the impedance Z_S to the impedance Z_{I_1}. The third, $2Z_{I_2}/(Z_R + Z_{I_2})$, is a similar transmission factor, indicating the modification due to the transition from Z_{I_2} to Z_R. The fourth, $e^{-\theta}$, gives the attenuation and phase shift due to a symmetrical network of transfer constant θ. The fifth, $\sqrt{Z_{I_1}/Z_{I_2}}$, indicates the transformer action of an asymmetrical network. The last, finally, is the so-called interaction factor, which arises from the multiple reflections in a network, connected into a circuit, where the impedances are not matched at either end.

If the transfer of power through the network is of primary interest, the two transmission factors and the interaction factor must be considered as introducing losses due to the lack of a good match at either end. But if the quantity of interest

is the output current or voltage, without any consideration of power, these factors may easily introduce gains rather than losses. For example, if the input impedance of the receiver is very high, as in the grid circuit of a pentode, the output current and power are very low, but the output voltage might become very large due to the transmission and interaction factors. This fact is of great importance to the transmission of interference, because it shows that it is by no means necessary to have an impedance match for maximum transmission of either voltage or current. On the contrary, a mismatch may actually introduce an increase in the interfering current or voltage at the output, or it may result in causing such large currents to flow from a power supply that the machine is damaged or circuits overloaded to the point where a fire hazard exists. Thus the practice of inserting large capacitors in circuits to reduce interference without careful consideration of what effects impedance mismatches may cause, is to be avoided. The reflection gain or loss in decibels resulting from a mismatch of impedances at a junction in a transmission line is given in Figure V-7 in Appendix V.

Similarly, the "transformer factor", $\sqrt{Z_{I1}/Z_{I2}}$, which produces the same modification of current and voltage as a perfect transformer, introduces neither gain nor loss, as far as the power is concerned. But, if the quantity of interest is current or voltage, this factor may cause an unexpectedly large gain. These facts explain why it happens that an interference source sometimes produces large disturbances in a particular receiver, even though laboratory measurements show no excessive currents or voltages when the source is tested alone.

1.5 APPLICABILITY OF THE "IMPEDANCE CONCEPT" TO THE SOLUTION OF INTERFERENCE PROBLEMS

The problem of combating interference is a threefold one: to minimize its generation, its transmission, and its undesirable effects on the receiver. The impedance concept is useful in all three of these, but particularly in the second. The transmitting network, which is the aircraft with all its equipment and wiring, is not designed with any specific purpose. In fact, it is not designed as a network at all. It is a haphazard conglomeration of wires and pieces of equipment which happen to form network elements. It is not surprising, then, that often a shield or filter, designed on the basis of considering only a small portion of this total network, turns out to increase, rather than decrease, the interference at the receiver. In almost all cases of this kind, the failure can be traced to the lack of consideration of the complete picture. By suppressing one mode of transmission, another might be sufficiently favored to produce the undesired result. A device designed to reduce the generation of interference might greatly increase its transmission, thus minimizing or even cancelling the effect of the reduction.

The tremendous increase in the importance of the radio interference problem in recent years is due to the following reasons:

(a) The large increase in the number of both receivers and potential interference sources in modern aircraft.

(b) The necessity of having a large number of pieces of electronic equipment crowded together in close quarters.

(c) The increased opportunity of transmission of interfering signals due to the large amount of wiring required in modern aircraft.

(d) The increased sensitivity of modern receivers.

(e) The increased use of higher and higher frequencies for communication purposes.

All but the last two of these causes are such that a unified overall approach, such as is suggested by the impedance concept, seems not only advantageous, but absolutely essential. One can hardly hope to cope with a problem that arises from the combined action of all the elements of the aircraft by isolating just a small part of it and designing a solution for that isolated part. A further discussion of the use of the impedance concept will be found in Appendix V.

1.6 METHODS OF TRANSMISSION

As far as the source is concerned, an interfering signal is sent into some impedance, which, in turn, may react back on the source to help determine the character of the signal. Whether or not the signal eventually reaches the receiver has no effect on the source. On the other hand, as far as the receiver is concerned, an interfering signal is received from an active network with a certain internal impedance. According to Thévenin's theorem*, it is only this impedance and the open-circuit voltage (or the short-circuit current) which need to be considered as far as the effect on the receiver is concerned. This section, however, is concerned with the way in which the interfering signal is transmitted from the source to the receiver, and, therefore, attention will be focused on the intervening network itself rather than on its input and output impedances.

In talking about networks (using this word in its most general sense), two points of view may be taken: viz., either the field or the circuit point of view. The field concept is by far the more general and is always applicable. The circuit concept is a special case, and, where applicable, leads to great simplifications. The ordinary circuit equations can be derived from Maxwell's equations, which describe the properties of the field under certain simplifying assumptions, the most important of which are the neglect of retardation effects, displacement currents, and phase variations of the current in a series circuit. All of these assumptions are excellent approximations at low frequencies, but become poorer as the frequency increases, and break down completely as the wave length becomes comparable in magnitude with the circuit dimensions. In some cases it is possible to account for high frequency effects by introducing artificial circuit parameters, such as the radiation resistance of an antenna, while still adhering to the circuit concept; but in general there will be a frequency above which the circuit concept must be abandoned. Certain low frequency phenomena, such as the coupling due to stray capacitances and

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*Thévenin's theorem states that, as far as the load is concerned, any two-terminal network containing a number of linear impedances and generators may be replaced by a single equivalent generator in series with a single impedance. The voltage of the equivalent generator is equal to the open-circuit voltage of the network, and the single impedance is equal to the impedance looking into the network with all the generators replaced by their internal impedances.

inductances, could also be treated better from the field point of view. However, since most engineers are more familiar with circuits, the circuit concept will be preferred in these cases. Nevertheless, the field concept will not be overlooked; and, whenever the circuit concept proves inadequate, the field concept will be utilized.

Two circuits are said to be coupled when currents or voltages in one produce corresponding currents or voltages in the other. According to circuit concepts, two circuits may be coupled either by a mutual impedance or by a mutual admittance. A mutual impedance exists between two circuits when the current flowing in circuit 1 produces a voltage in circuit 2. The magnitude of the mutual impedance is the ratio of the open circuit voltage of circuit 2, with all other voltage sources removed, to the current in circuit 1. A mutual admittance exists between a point of circuit 1 and a point of circuit 2 when the voltage between point 1 and some reference point (ground) produces a current to, or from, point 2, with this point connected to the same reference point. The magnitude of the mutual admittance is the ratio of the resulting current at point 2, to the voltage at point 1. Simple examples of mutual impedances are shown in Figure 1.6-A.

The mutual elements may be resistances, inductances, capacitances, or any series or parallel combination of these elements. Not all possible combinations, however, are equally important in the transmission of interference. When the coupling is of the mutual impedance type, the elements that must be considered most frequently are mutual inductance and the mutual impedance of a common ground return, which may arise, for example, from inadequate bonding. A good example of this is shown in Figure 1.6-B. Except for those cases of poor design or installation, a mutual resistance, capacitance, or self-inductance, i.e., any such element that is common to both the circuit of the interference source and that of the receiver, will not normally occur in a practical system since it would serve no useful purpose. This leaves the mutual inductance to be considered, which may very well be present, and, unfortunately, very often is. For, the presence of mutual inductance is nothing but a description, in circuit language, of the fact that the magnetic field set up by circuit 1 links with circuit 2, and therefore induces a voltage in it. <u>But a magnetic field exists around any circuit carrying current, and its linking with other circuits is very difficult to avoid.</u>

A somewhat similar situation exists when one considers the elements likely to become mutual admittances between two circuits. The ones most frequently encountered are mutual capacitances and the admittances offered by power or control cables connecting different pieces of equipment. Ideally, such cables should not carry any interfering currents, but when they do, the interfering signals are said to be transmitted by conduction. This method of transmission is important enough to be discussed separately in Paragraph 1.6.3, not only because such cables may constitute a direct metallic connection between an interference source and a receiver, but also because cables leading to receivers from entirely interference-free equipment may have interfering voltages induced in them, which are then conducted to the receiver. Moreover, a cable which is connected neither to an interference source nor a receiver, may serve as an intermediate path for an interfering signal which uses other methods of transmission before entering and after leaving that cable. Except for this, the only element of mutual admittance that might be present without serving any useful purpose as such is that of mutual capacitance. For, the

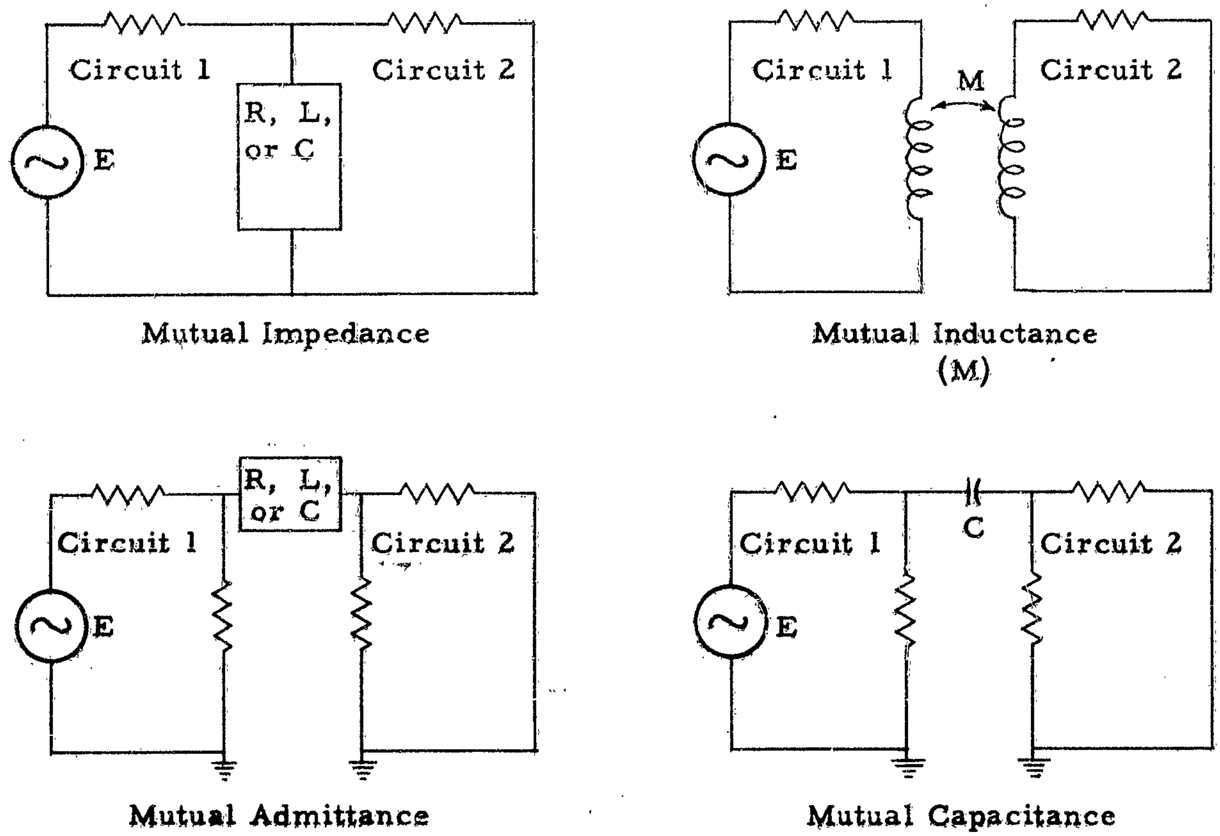


Fig. 1.6-A Coupled Circuits

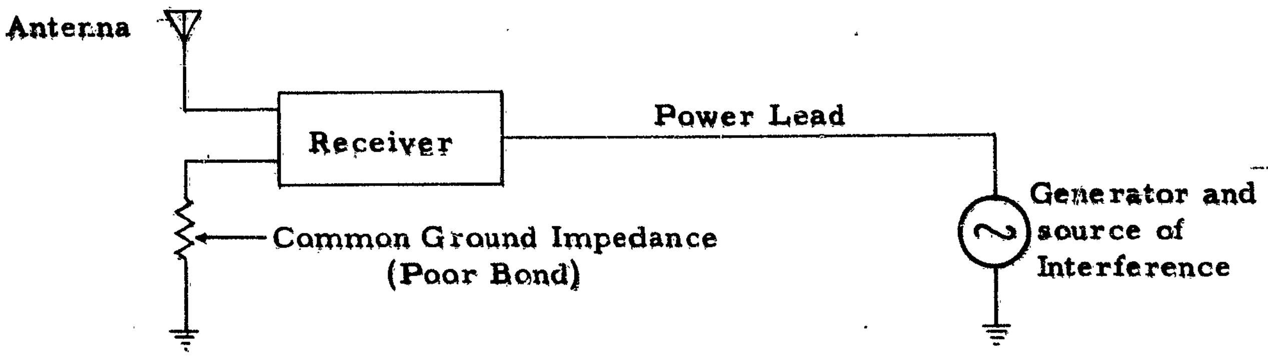


Fig. 1.6-B Mutual Impedance Coupling Through Common Ground Impedance

presence of mutual capacitance is nothing but a description, in circuit language, of the fact that an electric field exists in the vicinity of a conductor charged to some voltage, and that this electric field may induce a current in another circuit. Again, such interaction between two circuits is very difficult to avoid.

The effects of both mutual inductance, M, and mutual capacitance, C, increase with the frequency. This fact might seem contradictory if one considers that inductive reactance is $j\omega M$ and directly proportional to frequency, while capacitive reactance is $1/j\omega C$ and thus inversely proportional to frequency. The difference, however,

lies in the fact that the mutual inductance is considered as a mutual impedance, i.e., the current in the influencing circuit produces a voltage in the influenced circuit, while the mutual capacitance is considered as a mutual admittance, inasmuch as it is the voltage in the influencing circuit that produces a current in the influenced circuit. In the first case, the voltage produced is proportional to the frequency because $V = IZ$ and $Z = j\omega M$. It is true that M itself is not constant, but increases with frequency, but this does not affect the basic argument. In the second case, the current produced is also proportional to the frequency because $I = VY$, where the admittance $Y = j\omega C$. The same conclusions would be reached on the basis of an analysis starting with Maxwell's equations. In the case of coupling through a magnetic field, the law of induction states that the voltage induced is proportional to the rate of change of current, and thus, for a sinusoidally varying current, is proportional to its frequency. In the case of coupling through an electric field, the resulting current is the time rate of change of the accumulated charge on the conductors. Since the amount of charge is proportional to the voltage, the resulting current varies as the rate of change of the voltage, and, for a sinusoidally varying voltage, varies directly as its frequency.

The interfering signal may also reach the receiver by radiation. This mode of transmission cannot be treated by circuit methods because the phenomena leading to radiation were expressly neglected in the transition from field to circuit theory. Hence, this must be treated as a separate and distinct method of transmission as is done in Paragraph 1.6.4.

1.6.1 MUTUAL INDUCTANCE

In considering mutual inductance, it is important to keep in mind that mutual inductance exists between two complete circuits. To talk about the mutual inductance between the parts of two circuits, i.e., between two wires, each of which is a part of a separate complete circuit, has no meaning. By its very definition, mutual inductance involves a current in circuit 1 and a voltage in circuit 2, both of which must be complete circuits. The voltage in circuit 2 is measured between two end points after that circuit has been broken, so that the sum of all voltages induced anywhere in circuit 2 is measured. In other words, the action of an induced voltage cannot be localized. No answer can be given to the question, "What is the mutual inductance between two long straight wires?", because a long straight wire is not a complete circuit. It is true that many expressions for mutual inductance are listed in various handbooks for many different shapes and configurations of pairs of open wires, such as those given by F. E. Terman in the Radio Engineers' Handbook, pages 65-67. But these expressions have meaning only if they are used to compute the individual contributions from the component parts of two circuits, which must then be added together to obtain the mutual inductance between the two complete circuits. For example, the mutual inductance between two rectangular loops may be obtained by computing 16 individual mutual inductances (one for the mutual inductance between each side of one loop and each side of the other) and adding all of them together. Each of the quantities may be computed by means of the expressions found in the handbooks.

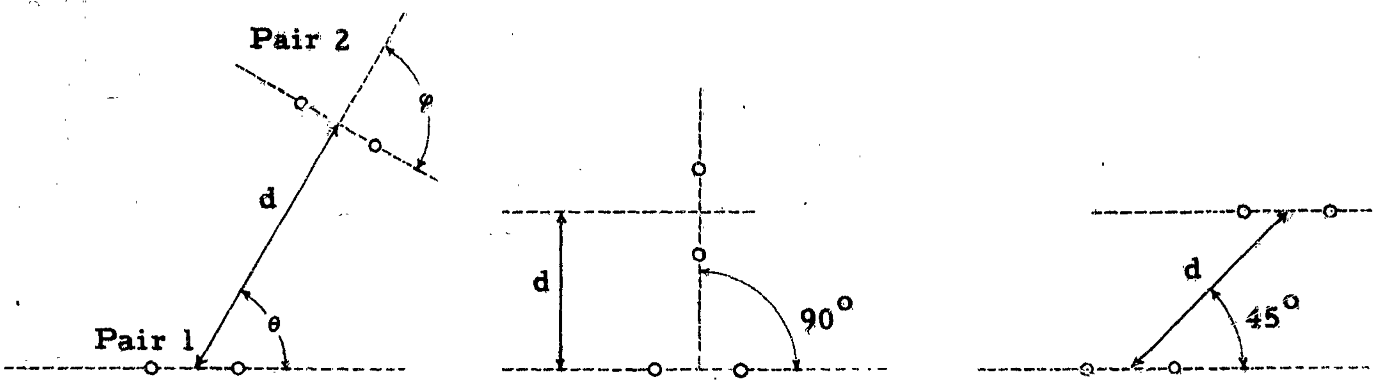
In general, the determination of the mutual inductance between two circuits of arbitrary geometry is a very difficult task. A case of great practical importance for the purposes of this book is that of two pairs of long straight wires all of which are parallel. A mathematical analysis of this configuration is carried out in Appendix VI.

The important results are given below.

The relative position of two pairs of long straight wires all of which are parallel may be specified in terms of the shortest distance, d, between the longitudinal axis of each pair, as shown in Figure 1.6.1 (A), and the two angles, θ and φ, which the plane of each pair makes with the separation d. Mathematical analysis shows that, for values of d large as compared to the spacing between the wires of each pair, the mutual inductance varies inversely as the square of the quantity d. When d is small, the mutual inductance starts at a finite value and decreases at a much slower rate than would be given by the inverse square law. For every value of θ, there is one value of φ between 0 and 180°, such that the mutual inductance is zero, and another value of φ, 90° removed from the first, such that the mutual inductance is a maximum. When d is large as compared to the spacing, the value for zero coupling is given by:

$$\varphi = 90^\circ - \theta \tag{13}$$

This leads to the interesting result that, when d is in the plane of, or perpendicular to the plane of one pair, there is zero coupling when the planes of the two pairs are perpendicular to each other. But if d makes an angle of 45° with the plane of one, the coupling is zero if the planes of the two pairs are parallel. This is shown in Figures 1.6.1, (B) and (C). The situation is reversed if the coupling is to be a maximum.



(A) Notation (B) (C) Two Positions of Zero Coupling

Fig. 1.6.1 Mutual Inductance Between Two Pairs of Parallel Wires

This analysis was carried out for a rather special case. Two conclusions, however, can be expected to remain valid for other cases. First, for a pair of two-wire systems, the mutual inductance varies initially in some complicated way with the distance between the two circuits, but as the distance between them becomes large as compared to the dimensions of each, it will vary inversely as the square of the distance. This also applies to single-wire systems with ground returns because these may be reduced to two-wire systems by the method of images. This law, however, might be modified considerably by the presence of other conductors in the vicinity. Secondly, no matter how complicated the geometry of the circuits, for a given distance between centers, there will always be a position of circuit 2 relative to circuit 1 such that the coupling is zero. This last statement is obvious if one considers that the net magnetic flux must pass through circuit 2 in one of two directions, and that this direction can be reversed by reversing the position of circuit 2. In going from one position to the other, the flux changes continuously from

maximum positive to maximum negative, and in doing so it must of necessity pass through zero for some position of circuit 2.

1.6.2 MUTUAL CAPACITANCE

While, according to the principle of duality, certain similarities exist between the effect of capacitive coupling and that of inductive coupling, there are also a number of significant differences. It is impossible to speak of mutual inductance except between closed circuits. On the other hand, the definition of mutual admittance introduces the concept of the mutual capacitance between just two points of two circuits. It is not very practical to use this concept because in all physical problems there will be capacitances between many points or regions of the two circuits. But for purposes of analysis, it does make sense, and it is much simpler, to use only one point at a time in each circuit. In magnetically coupled circuits, the induced voltage can be reversed by proper placement of one circuit with respect to the other. From this it could be concluded that there must be a position of zero coupling. On the other hand, in capacitively coupled circuits the direction of the current flowing to or from a point in one circuit depends only on the polarity of the voltage in the other. Therefore, it cannot be reversed simply by manipulation of the relative positions of the two circuits. From this it may be expected, and analysis shows it to be true, that capacitive coupling cannot be reduced to zero by rotation of one circuit about its center.

If the case of two long straight parallel wires is analyzed, it is found that the capacitance between these wires varies inversely as the logarithm of the distance between them. This would indicate that, for large distances of separation between two parallel wires, the capacitance between them decreases much more slowly with distance than the mutual inductance. This conclusion, however, cannot be generalized for other systems, because the law of variation with distance for capacitance depends very much on the shape of the wires and their relative positions to other metallic objects in the vicinity. A general statement which can be made is that the capacitance between two points in two circuits decreases with distance somewhat more slowly than according to the inverse square law.

The definition of mutual admittance indicates that two circuits involved must have a common ground connection in order for a current to flow through the mutual element. It must not be concluded that the effects of mutual admittance may be prevented by eliminating all ground connections from one circuit. Practically, isolation is quite impossible without complete shielding (which would reduce the mutual admittance itself to zero) because even in the absence of a metallic connection to a common ground, there is always capacitance to some metallic object, which may be the airframe, providing a return path for the radio frequency current through the mutual element.

1.6.3 CONDUCTION

Whenever there is a direct metallic connection between two circuits, and in addition a return path, a conduction current may flow between these two circuits. The return path may be another metallic lead, or a mutual capacitance, or most frequently the metallic structure of the aircraft acting as the ground return. The magnitude of the resulting current depends both on the potential difference between

the points of exit and entry in the exciting circuit, and on the total loop impedance between the same points. It is important to remember to include the impedance of the return path as well as the impedance of the direct connection when computing this total loop impedance. If all impedances involved are linear, the current may be computed by a simple application of Ohm's law, $i = E/Z$, in accordance with the circuit concept.

The most common example of this is the transmission of interfering signals through the power and control leads, both out of the interference generators and into the receivers. The circuit shown in Figure 1.6.3 illustrates how an interfering signal may be transmitted from a motor into a receiver if both are connected to the same power supply.

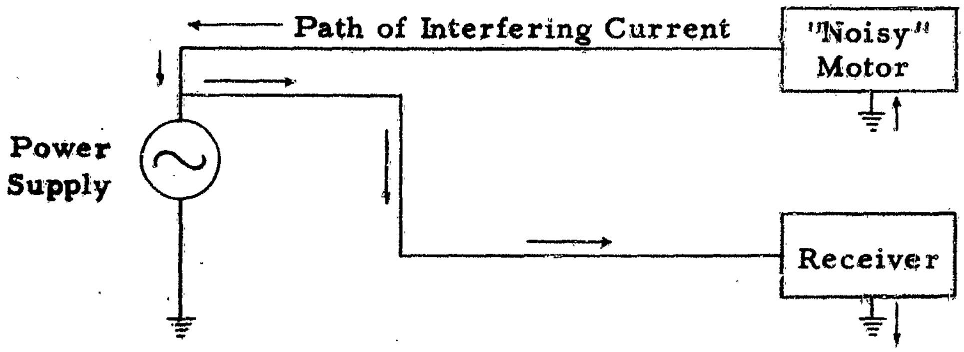


Fig. 1.6.3 Transmission of an Interfering Current by Direct Conduction

The field concept is just as applicable to the case of conduction as to the cases of mutual inductance and capacitance discussed above. Strictly speaking, the energy associated with the interfering current is transmitted not through the metallic leads, but through the fields surrounding these leads, the leads acting only as guides. Yet it is precisely here that the circuit concept leads to the greatest simplifications and is, therefore, most frequently used. But in the discussion of radio interference it is often very useful to keep the field concept in mind, because it draws attention to the fact that there are always strong interfering fields in the vicinity of conductors carrying interfering currents. In fact, many troublesome interference problems are caused by bundling power or control cables leading to a receiver together with leads carrying interfering currents, thus exposing them to strong interfering fields.

As an example of a case that should be treated by the application of the field concept, rather than the circuit concept, consider a cable connector with an unused pin. The pin does not actually connect two circuits, yet it may carry interfering conduction currents by having one end act as a receiving antenna and the other end as a transmitting antenna. Here the pin should be considered as a wave guide which transmits an electromagnetic wave from one side of the connector to the other.

1.6.4 RADIATION

The term "radiation", technically speaking, is used to describe the phenomenon of electromagnetic waves spreading out in space from a source, according to the laws of wave propagation. Nevertheless, the term "radiated noise" has become so commonly used to mean any interfering signal detected through the medium of an

electric or magnetic field, that it would not be possible to restrict the words "radiation" and "radiated" to their technical meaning without rewriting much of the literature on radio interference and almost all of the specifications on interference measurements. It is clear that the transmission of interfering signals through mutual inductances and mutual capacitances, as discussed in Paragraphs 1.6.1 and 1.6.2, would be classified as radiated interference in the loose sense of the word.

Examples of true radiation in the sense defined above, within the aircraft, are extremely rare. An example of true radiation outside the aircraft would be the transmission of an interfering signal from a transmitting antenna to a receiving antenna mounted several wave lengths away. When the receiving antenna picks up an interfering signal which is apparently radiated from a nearby opening in the skin of the aircraft, the transmission is usually not through true radiation but through capacitive coupling from the antenna to the outside surface of the aircraft, which has acquired a charge due to currents flowing around the edges of the opening.

Radiation in the strict sense is not amenable to treatment with the methods of ordinary circuit analysis because radiation is caused entirely by the effects of retardation, the phenomenon specifically neglected in the transition from field to circuit concepts. If an electromagnetic disturbance could be propagated in space with infinite velocity, the fields that give rise to the effects of mutual inductance and capacitance would be unchanged, but the radiation field would be absent, a fact that can easily be demonstrated mathematically.

The mathematical expression for the total fields in the vicinity of a conductor carrying time-varying currents is very complicated and cannot, in general, be found analytically, except in very simple cases. The cases where it is possible to find such expressions are the ones involving very simple geometric arrangements, such as a thin straight wire or a thin circular loop. It can be shown, however, that the mathematical expression for the field due to any source can be developed into a series of terms, each of which represents the field due to certain simple arrangements of conductors. The first term in this series gives the field due to an electric dipole; the second, that due to a magnetic dipole and an electric quadrupole, etc. This expansion is useful only if the mathematical series converges rapidly, so that the first few terms give a good approximation to the total field. The rapidity of convergence depends on the ratio of r, the distance of the point of observation to the center of the source, to a, the radius of the smallest sphere that can be placed so as to enclose all parts of the source. (See Figure 1.6.4-A.) When this ratio is very large, only the first term in the expansion is important (unless this term happens to be zero, in which case only the first non-vanishing term is important). The result is dipole radiation. When this ratio is less than unity, the expansion does not converge at all. This possibility of "multipole expansion" explains why the consideration of very simple, but non-practical cases nevertheless has much significance for the analysis of the more complicated, practical cases.

The two simplest cases are those of an oscillating electric and magnetic dipole. The electric dipole is represented physically by a very short length of very thin wire carrying a uniform current that varies sinusoidally in time. The magnetic dipole is represented physically by a very small circular loop carrying a uniform current that varies sinusoidally in time. For the electric dipole the components of the fields surrounding the conductor of length L are given in spherical coordinates by the

following expressions, assuming $L > r$ and $L \ll \lambda$:

$$E_\theta = \frac{L I \sin\theta}{4\pi\varepsilon}\left[\frac{\sin\omega(t-r/v)}{r^3\omega} + \frac{\cos\omega(t-r/v)}{r^2 v} - \frac{\omega\sin\omega(t-r/v)}{r v^2}\right] \quad (14)$$

$$E_\varphi = 0 \quad (15)$$

$$E_r = \frac{2 L I \cos\theta}{4\pi\varepsilon}\left[\frac{\sin\omega(t-r/v)}{r^3\omega} + \frac{\cos\omega(t-r/v)}{r^2 v}\right] \quad (16)$$

$$H_\theta = 0 \quad (17)$$

$$H_\varphi = \frac{L I \sin\theta}{4\pi}\left[\frac{\cos\omega(t-r/v)}{r^2} - \frac{\omega\sin\omega(t-r/v)}{r v}\right] \quad (18)$$

$$H_r = 0 \quad (19)$$

where $v = 3 \times 10^8$ meters per second is the velocity of electromagnetic waves in free space, $\varepsilon = 8.854 \times 10^{-12}$ farads per meter is the absolute permittivity of free space, and the other symbols are explained in Figure 1.6.4-B. In the case of the magnetic dipole, the components of the fields surrounding the small loop of area A and carrying a current $I \cos \omega t$, assumed to be at the origin in the x-y plane, are given by the following expressions:

$$E_\theta = 0 \quad (20)$$

$$E_\varphi = \frac{IA\sin\theta}{4\pi}\,\omega\mu\left[\frac{\sin\omega(t-r/v)}{r^2} + \frac{\omega\cos\omega(t-r/v)}{r v}\right] \quad (21)$$

$$E_r = 0 \quad (22)$$

$$H_\theta = \frac{IA\sin\theta}{4\pi}\,\omega\left[\frac{\cos\omega(t-r/v)}{r^3\omega} + \frac{\sin\omega(t-r/v)}{r^2 v} - \frac{\omega\cos\omega(t-r/v)}{r v^2}\right] \quad (23)$$

$$H_\varphi = 0 \quad (24)$$

$$H_r = \frac{2\,IA\cos\theta}{4\pi}\,\omega\left[\frac{\cos\omega(t-r/v)}{r^3\omega} - \frac{\sin\omega(t-r/v)}{r^2 v}\right] \quad (25)$$

It is seen that in the above equations there are three kinds of terms, which differ in their variation with r. The terms that vary as $1/r^3$ represent an electrostatic dipole. The terms that vary as $1/r^2$ represent what is called the induction

field. Finally, the terms that vary as $1/r$ represent what is called the radiation field. The relative importance of the various kinds of terms depends on the ratio of r to λ, where $\lambda = 2\pi v/\omega$ is the wave length of the radiation. When r/λ is much smaller than unity, the radiation terms are negligible. When r/λ is much larger than unity, the static and induction terms are negligible. When $r = \lambda/6$, approximately, the induction and radiation fields are equal.

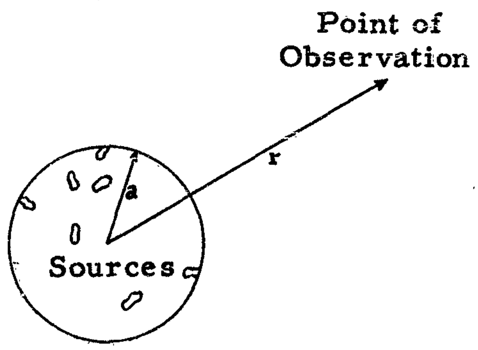


Fig. 1.6.4-A Quantities of Length in Radiation Problems

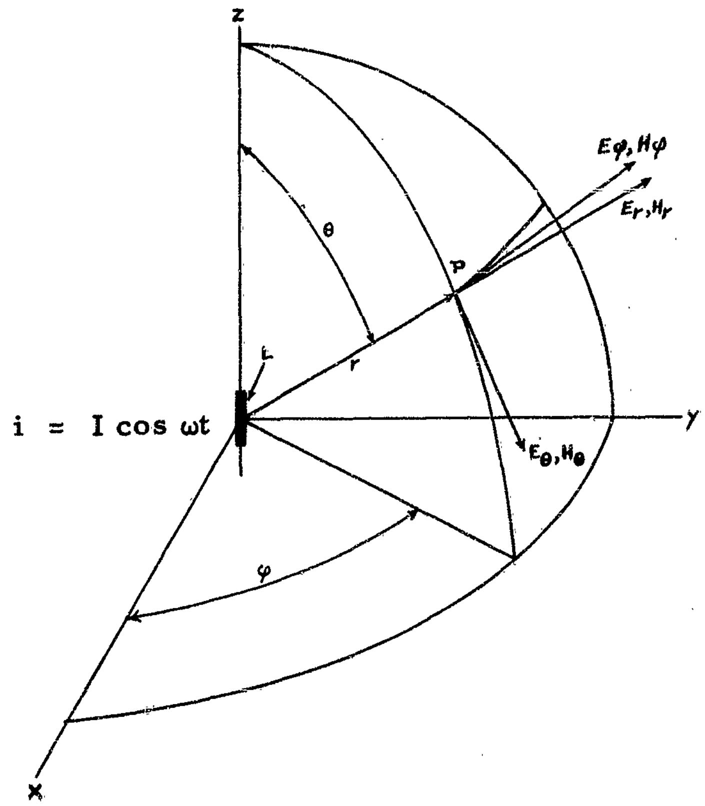


Fig. 1.6.4-B Dipole Radiation

Comparison of the case of the magnetic dipole (current loop) with that of the electric dipole shows that the structures of the two fields are the same except that the poles of the electric and magnetic fields are reversed.

Based on the structure of the electromagnetic field in the vicinity of an oscillating dipole, the statement is sometimes made that for any radiating system there is an induction field, varying as $1/r^2$, and a radiation field, varying as $1/r$. This statement, however, is misleading. Although it may be true if properly applied, generally it is false when applied to most problems arising in the propagation of interfering signals within an airplane.

To determine the correct applications of this statement, a clear distinction must be made between the three quantities of length entering every problem of this kind, illustrated in Figure 1.6.4-A. These are (1) r, the distance to the point of observation from the center of the source; (2) a, the radius of the smallest sphere that can enclose all of the source, and (3) λ, the wave length. This last quantity, the wave length, does not have a unique meaning unless the time variation is sinusoidal. For some other variation, say an interference pulse, each frequency component must be considered separately.

The point that is frequently forgotten is that the validity of the statement made above depends entirely on the relation between r and a, not r and λ. The statement holds whenever r is $\gg$ a, and the configuration of the source is such that there is dipole radiation. In practice, neither condition is ever satisfied within the aircraft except, possibly, at ultra-high and super-high frequencies. The first condition fails because the dimensions of the potential radiators (the aircraft structure itself is an important example) are usually of the same order of magnitude as the distances between the radiator and the receiver or pick-up lead, due to the proximity of electrical equipment in aircraft. The second condition fails because there are present within the aircraft so many absorbers, reflectors, and systems capable of re-radiation, that the resultant field at any point will consist of several waves of different amplitudes and phases superimposed, and the laws of simple dipole radiation would never apply.

It must also be remembered that the separation of the total field into an induction and a radiation field is by no means physical, but entirely analytical. Any possible physical measurement will always yield a measure of the total field, not its analytical components. When the basic conditions mentioned above are not fulfilled, even the analytical separation is no longer possible. The total field must then be considered and its law of variation with distance will, in general, be very complicated. It cannot even be said that the field will always decrease as the distance from the prime radiator increases. Because of the possibilities of reflection and re-radiation, standing waves may be set up and the field may actually increase in certain regions, as the distance from the prime source is increased, a condition which is called resonance excitation.

In general, the conditions in aircraft are almost always such that the fields of interest are those observed very close to the source. In the frequency range from about 30 to 300 megacycles per second, the distances of interest are of the same order of magnitude as a wave length. Even if there were true dipole radiation, no approximations could be made in this region and the total field would have to be considered. The actual radiation is likely to be much more complicated than dipole radiation, and such simple approximations will be even less valid.

True radiation is more likely to be important at frequencies above 300 megacycles per second. The rather thorough treatment of radiation in this paragraph is justified by the increased use of the ultra-high frequency and super-high frequency regions in modern aircraft. Moreover, most of the above statements about the relative unimportance of true radiation were based on the fact that, in order for any body to be an efficient radiator, its dimensions must be at least of the order of magnitude of a wave length. But because of the extremely high sensitivity of modern receivers, even a very inefficient radiator could produce a troublesome radiation field at the receiver.

In considering the interfering fields both close to and far from the source, it is sometimes important to know the significance of the electric field, relative to that of the magnetic field. Confusion may arise from the fact that, according to Maxwell's equations: a varying electric field is always accompanied by a magnetic field; a varying magnetic field is always accompanied by an electric field; and the ratio of the two, called the impedance of the medium, is a constant (377 ohms for air) for a plane wave. A careless interpretation of these facts leads to the conclusion that it

makes no difference which is considered for all possible conditions, since the two always occur together and bear a fixed ratio to each other. This reasoning is in error, because what exists within the aircraft is usually not a plane wave. The impedance for more complicated types of waves, such as cylindrical or spherical waves or combinations of these, is not, in general, equal to 377 ohms. It may take on almost any value, and often varies rapidly from point to point in the vicinity of the source. A very small impedance means that the electric field is small as compared to the magnetic field. A very large impedance means that the magnetic field is small as compared to the electric field. Since some receivers are more sensitive to one type of field than the other, a knowledge of the impedance would be of great practical value in many cases. Unfortunately, the evaluation of the impedance is practically impossible for the complicated field configurations encountered in the interior of modern aircraft.

1.7 SUSCEPTIBILITY OF RECEIVERS

If the interfering signal is not suppressed at the source and if its transmission is not prevented, it will then reach the receiver. The basic ways in which a signal may enter the receiver are:

(a) through the antenna, antenna lead-in, or antenna receptacle,
(b) through the power or control leads,
(c) through the output leads connecting the receiver to earphones, intercommunication systems, scopes, or other indicating devices, and
(d) through the casing enclosing the receiver, or any joints or openings in the casing.

The effect on the receiver will be greatest if the interfering signal enters in the same way as the desired signal, because then it will immediately act on the most sensitive part of the receiver and may mix with the desired signal to such an extent that the separation of the two signals becomes practically impossible. For most receivers this most dangerous point of entry is given in (a) above. In others, such as a remotely controlled actuator, an interfering signal may enter through the control lead to produce a false actuating pulse.

If the interfering signal enters in a way different from that of the desired signal, the path of entry becomes just another link in the transmission system through which the interfering signal is moving. Eventually, it must enter the same path as the desired signal and mix with it. Otherwise it cannot reach the output stage of the receiver and cause undesired response or malfunctioning. Thus, if it enters through the power lead, it is transmitted into the receiver by conduction and may then enter the path of the desired signal, say, through inductive coupling of a power transformer with a radio frequency coil, or through capacitive coupling from the filament to the grid of an amplifier vacuum tube. If it enters through the headphone leads, it is again transmitted into the receiver by conduction and may then couple, within the receiver, with a sensitive circuit. If it enters through the casing or any openings in the casing, it is transmitted usually by capacitive coupling. Occasionally, it may be transmitted by inductive coupling as, for example, when an interfering field outside induces currents in the casing, which, in turn, produce an interfering magnetic field inside, or, in rare cases, even by radiation. In all these cases, the problems encountered are truly problems of transmission as treated in Paragraph 1.6. One

half of the proper design of receivers for interference-free operation is simply design for minimum transmission of interference into the receiver.

The remaining portion of this subsection will deal with the effect on the receiver of the interfering signals which have, in one way or another, entered the path of the desired signal in the receiver.

1.7.1 THE NUISANCE VALUE OF INTERFERING SIGNALS

The final effect of any interfering signal is to make it difficult or impossible for the receiver to function properly. The extent to which an interfering signal adversely affects the proper functioning of the receiver is called its nuisance value. Depending on the type of receiver, the effect may be complete inability to receive a message, or false indications of navigation instruments, or some such drastic "malfunctioning" as the premature explosion of a warhead triggered by a spurious control impulse. At any rate, in order for the interfering signal to have a nuisance value, either it must eventually contain one or more frequencies within the normal output range of the receiver, or it must be capable of preventing the receiver, or at least one of its stages, from functioning properly. In other words, the interfering signal must be capable of either producing effects similar to the desired signal or of preventing the desired signal from having its normal effects.

Normally, to produce these effects at the output of the receiver, the interfering signal must contain frequencies within the acceptance band of the receiver. But it must be remembered that the acceptance band of a receiver may be much larger for interfering signals, which may be of any magnitude, than for desired signals, which normally do not exceed a certain level. Most receivers have input circuits that behave somewhat like bandpass filters. Ideally the response to frequencies outside the passband is nil, but practically the attenuation in the attenuation bands is never infinite. Even though it will completely suppress all signals, outside the pass band, that are of the same order of magnitude as the desired signal, it may still permit very strong interfering signals which lie entirely outside the pass band to enter and to be transmitted past the first input stage. Once they have gone as far as that, they may produce one or more of the following three effects: (1) they may be strong enough to be transmitted directly to the output in spite of further attenuation in succeeding selective stages such as the intermediate frequency stages of superheterodyne receivers, (2) they may be strong enough to overload one or more stages thus making the receiver inoperative, and (3) they may combine with other signals in a non-linear element in such a way as to produce a new frequency that is within the band of accaptance of the receiver. For example, they could combine in the mixer tube with a harmonic of the local oscillator in a superheterodyne receiver to produce a signal of the intermediate frequency. This last effect is sometimes incorrectly called "cross-modulation", a term that properly applies to the more complicated effect of a carrier being modulated by the modulating frequencies of another carrier nearby.

When the final output of the receiver is an aural or visual signal to be interpreted by a human operator, it is extremely difficult to assign any quantitative measure to the nuisance value of an interfering signal. In the final analysis, it is the operator who must decide what the real nuisance value of such a signal is, under the worst possible flight conditions and the strain of long combat missions. Certain general statements may, however, be made. If, in an audio receiver, the frequency

spectrum of the interfering signal is fairly evenly distributed over the entire audio range, the result is usually considered to have a very "high nuisance value" by most operators. This is the case when the output consists of the crashes and background rumbles characteristic of atmospheric disturbances, or the pops and cracking sounds found in ignition interference, or the hash sounds usually accompanying interference from commutators. If, on the other hand, the interfering signal contains only one, or a very few, frequencies, the resulting steady note of definite pitch may not be very bothersome at all to some operators unless it is extremely loud. For example, a 60 or 400 cycle power line hum of moderate strength, while unpleasant, may not impair the intelligibility of the desired signal to an appreciable extent. However, even this type of interference must be avoided because two or more "moderate" interfering signals may easily combine to produce a level of interference sufficient to drown out all desired signals. In general, the nuisance value of an interfering signal increases with the number of frequencies contained in it, but the effect of each frequency depends on the sensitivity of the individual operator. This sensitivity usually increases gradually for the lower frequencies, reaches one or more maxima somewhere between 700 and 3000 cycles per second, and then decreases to zero at about 15,000 cycles per second. Individual differences between operators are not quite so pronounced when the receiver is a radar scope; but here too, the skill and experience of the operator play an important part in determining the nuisance value of an interfering signal during actual operations.

1.7.2 PULSE LENGTHENING

Any interfering signal that enters the receiver through the antenna or antenna lead-in, must pass through the same stages of the receiver as the desired signal, before it assumes its final form. During its passage through these stages, it may be modified both linearly and non-linearly. It experiences non-linear distortion whenever the output of a stage contains frequencies that were not present in the input. The appearance of intermediate frequencies in the mixer stage of a superheterodyne receiver is an important example of this. Linear distortion occurs whenever different frequencies experience different attenuations and different equivalent velocities of propagation. It is here that the difference between the desired and interfering signals is usually most important.

The input stages of a receiver are designed to pass the desired signals with a minimum of distortion of any kind. The desired signal usually contains frequencies within a definite range, say, a carrier and two side bands in amplitude modulation, or a carrier and about 20 to 30 sidebands in frequency modulation. A well-designed receiver will pass a signal whose spectrum is confined entirely to its band of acceptance with little or no distortion. On the other hand, a signal whose spectrum extends considerably above and below its acceptance band may be modified beyond recognition by the same receiver. Most interfering signals extend over a frequency spectrum much larger than the acceptance band of the receiver. Therefore, the wave form of the interfering signals will be considerably modified in its passage through the receiver.

The most frequent and most important effect of this kind is the lengthening of interference pulses commonly referred to as "shock excitation" or "ringing". It occurs whenever a pulse of short duration enters a circuit whose bandwidth is narrower than the frequency spectrum of the pulse. In that case, the pulse at the output

will be much longer than that at the input and considerably modified in shape. If the circuit has the characteristics of a resonant circuit, and if the pulses occur comparatively far apart, the output pulse will consist of a damped oscillation (see Figure 1.2-B) at the resonant frequency. The circuit has been excited by the "shock" of the pulse and, as a result, "rings" at its resonant frequency. But even if the characteristics of a resonant circuit are absent, the output pulse will be much longer than the input pulse. The only circuit that will pass a short pulse without distortion is a circuit without inductance or capacitance, or in other words, a circuit whose acceptance band is infinite.

An example of the lengthening of a short pulse in various stages of a superheterodyne receiver is shown in Figure 1.7.2. Part (A) of this figure shows an approximately rectangular pulse of 0.4 microsecond duration, which is applied to the input of the receiver. The next part (B) shows the resultant "ringing" of the radio frequency stage. Figure 1.7.2, (C) shows the further lengthening of the pulse in the intermediate frequency stage. Finally, Figure 1.7.2, (D) shows the pulse as it appears at the audio output after rectification. Note that the pulse is now almost 200 microseconds long, i.e., the initial duration has been multiplied by a factor of almost 500.

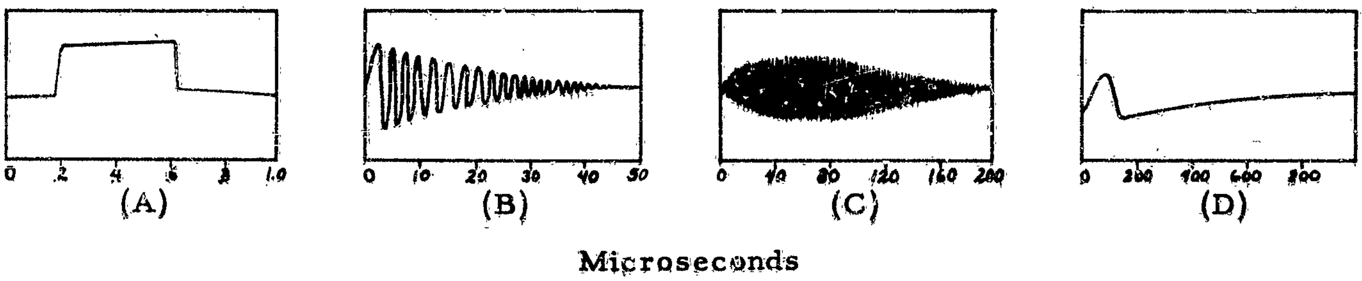


Fig. 1.7.2 Pulse Lengthening in Various Stages of a Receiver

There is a definite relationship between the amount of pulse lengthening and the bandwidth of a stage. Roughly, the lengthening of a pulse is inversely proportional to the bandwidth of the stage through which it passes. This rule is valid both for unidirectional rectangular pulses and for high frequency sinusoidal pulses produced by amplitude modulation of a high frequency carrier with a rectangular modulation envelope.

1.7.3 OVERLOADING

Another effect may be caused by interfering signals in a receiver. If the amplitude of a signal is much larger than that for which a stage is designed, it may completely block that stage so that no useful signal can get through at all. Whether this limiting action is intentional or not makes little difference. The fact that the stage is inoperative for the duration of the high amplitude signal will always cause distortion in the desired signal. This effect is often produced intentionally because the resulting distortion may be less than that which would occur if the interfering signal were allowed to pass through unlimited.

The most common cause of overloading is the saturation of a vacuum tube. A tube is a linear device only for a certain range of applied voltages and currents. If

an excessive signal is applied to the grid of a triode, the plate current may become temperature limited, which means that the output becomes independent of the input, and dependent only on the cathode temperature and surface condition. The same effect may occur in a diode that is called on to rectify an excessive signal.

1.7.4 EFFECTS OF DIFFERENT TYPES OF MODULATION

Whenever radio frequencies are employed for the transmission of intelligence, use is made of some type of modulation. The original signal is not transmitted directly, but rather it is used to modulate a so-called carrier, whose frequency is much higher than that of the original signal This is done both in order to increase the number of available communication channels, and because low frequencies (below about 100 kc) cannot be efficiently transmitted by radiation. The most common types of modulation are amplitude, frequency and phase, and pulse modulation.

The three types of modulation differ markedly in their abilities to transmit useful information in the presence of interfering signals. This ability consists of two different and entirely separate properties. One is the ability of the receiver to amplify the desired signal more than the interfering one. It is measured in terms of an interference-voltage reduction factor, defined as the ratio of the signal-to-interference ratio at the output to that at the input. The second is the ability to separate the desired signal from the interfering one, which is measured in terms of an improvement threshold, defined as the minimum signal-to-interference ratio necessary at the input to produce an intelligible signal at the output. The signal-to-interference ratio required at the output for intelligible reception depends on the type of signal, i.e., whether the intelligence being received is voice, code, or a control signal, as well as on the type of interference that is present. The improvement threshold may be thought of as the point at which the desired signal "takes over" so that it, rather than the interference, determines the main character of the output.

To explain the practical significance of these two properties, consider a receiver to which a combination of signal and interference is applied with ever-increasing signal-to-interference ratio. At first the interference is much larger than the desired signal and the latter is completely masked at the output. No useful information is received. As the desired signal increases in strength, a point will be reached at which the desired output signal becomes intelligible. The input signal-to-interference ratio at this point is the improvement threshold. At and beyond this point, the amount by which the ratio is reduced between the input and output is measured by the interference-voltage reduction factor.

The behavior of the three types of modulation with respect to these two properties is illustrated in Figure 1.7.4, in which the signal-to-interference ratio at the output is shown as a function of the signal-to-interference ratio at the input for some typical cases. The improvement threshold is the point of maximum slope on each curve, i.e., the point at which the desired signal "overtakes" the interference at the most rapid rate. The interference-voltage reduction factor in each case is given by the ratio of ordinate to abscissa. The values of this ratio of greatest interest to the designer are those for which the signal-to-interference ratio at the input is in the region of 10-20 db.

These characteristics of the various types of modulation are important for

equipment designers to consider in the initial design stages especially when there is a choice as to what type modulation is to be used. Once the modulation is decided upon, the designer must make use of the best design techniques in order to insure that the system is interference-free within the limitations posed by the type of modulation.

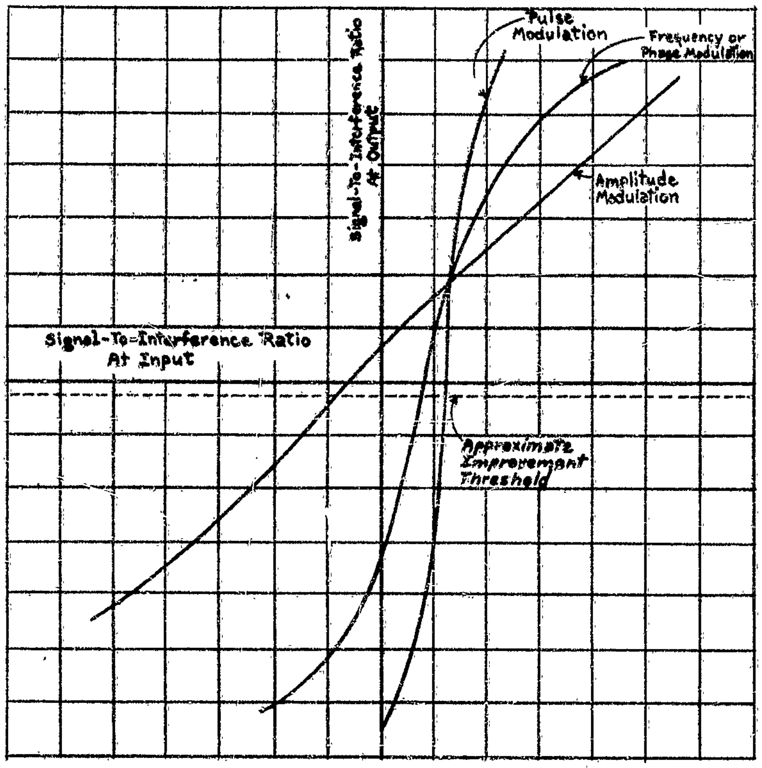


Fig. 1.7.4 Relationship of Signal-to-Interference Ratio at Output to that at Input for Different Types of Modulation

Another consideration is important to the overall picture for interference-free operation of the complete aircraft. It is the fact that all three types of modulation may be involved in the various equipments installed. Therefore, for interference-free operation of the complete aircraft, techniques must be applied as required for satisfactory operation of the most interference susceptible equipments.

1.8 BASIC METHODS FOR SUPPRESSION OF RADIO INTERFERENCE

Basically there are three ways of combating an interfering signal: To minimize (1) <u>its generation</u>, (2) <u>its transmission</u>, and (3) <u>its undesirable effect on the receiver.</u>

To prevent the generation of interfering signals means to design all equipment in such a manner that no interference will be generated. This is obviously the ideal way of dealing with interference problems because, if no interference were generated, no further attention would have to be paid to the problem. In some cases, this can be achieved simply by a proper choice of components. For example, if there is a choice between two motors for a particular application, one with a commutator and

one without, then, all other things being equal, the one without a commutator should be chosen because this choice would eliminate all interference due to commutation. Or, in making a choice of a device to obtain alternating current from a direct current supply, vibrators should be avoided, other considerations being the same, because the arcing that accompanies their operation is a more serious source of interference than that associated with other types of inverters.

If the generation of interfering signals cannot be prevented entirely, as, for example, in those sources for which the generation of interference is inherent in their normal functions, good design may be able to minimize it. For example, in the ignition system of the aircraft engine, the spark is essential and cannot be eliminated. But the exceedingly steep wave fronts, which usually accompany the spark and are the primary cause of the strong interference fields, may not always be essential to the proper operation of the ignition system. Here proper design may be able to prevent some of the interfering signals from being generated though it can never eliminate all of them.

The prevention or minimizing of the generation of interfering signals is properly the task of the designer of components of aircraft electrical systems though some prevention can also be achieved by the aircraft manufacturer in the assembly of systems. The application of the basic principles developed in Paragraph 1.8.1 to the design of components will be found in Paragraph 3.2.

Since it is never possible to prevent all interfering signals from being generated, the designer must try to keep them from reaching the receiver. This problem may be attacked at various points. The task of preventing the interfering signals from leaving the immediate vicinity of the source also belongs properly to the designer of components. For example, interference from a motor may be kept "bottled up" entirely within a metal housing, which is an integral part of the motor and, if carefully designed, may act as a complete shield. The task of preventing the transmission of interfering signals at points between components belongs to the designer of aircraft systems and the engineer responsible for the installation of components and systems within the aircraft. The application of the basic principles developed in Paragraph 1.8.2 to the design and layout of aircraft systems will be found in Paragraph 3.3.

Finally, an effort must be made to prevent the interfering signals from affecting the receiver. Partly, this is a matter of preventing its transmission into the receiver or its reaching any sensitive circuits within the receiver as pointed out in Paragraph 1.7. Partly, this is a matter of incorporating special circuits in the receiver which, in one way or another, reduce its nuisance value as defined in Paragraph 1.7.1. The task of minimizing the effect of interference on the receiver belongs properly to the designer of receivers. The application of the basic principles developed in Paragraph 1.8.3 to the design of aircraft receivers will be found in Paragraph 3.4.

1.8.1 SUPPRESSION AT THE SOURCE

While there is some question as to the best point of application of interference suppression techniques in commercial aircraft, there can be no doubt that for <u>military aircraft, with their tremendously complex arrangement of electrical equipment</u>

and with the ever present possibility of addition of new or relocation of old equipment, suppression at the source must never be neglected. For this reason, it is highly desirable to have the interference-suppressing features "designed into" the equipment for military aircraft before their delivery to the aircraft manufacturer. It cannot be overemphasized that the designer must always keep this in mind while he is designing the equipment to meet its functional characteristics.

Suppression at the source has two basically different aspects: The prevention or minimizing of the generation of interfering signals, which is treated in this paragraph, and the "bottling up" of the generated interfering signal within the source, which is a problem of transmission and will be treated in Paragraph 1.8.2.

In Paragraph 1.3, the origin of interfering signals was traced to variations of either generated electromotive forces or impedances. According to this, preventing or minimizing the generation of interference consists of preventing or minimizing all unnecessary variations of electromotive forces and impedances. This means that, for each piece of equipment, the designer must analyze the causes of variations of electromotive forces and impedances, eliminate those that are unessential to the proper operation, and reduce the essential ones to the absolute minimum. Further consideration of this aspect will be illustrated by the analysis of individual components in Section 3. Two cases, however, involve considerations of a general type and will be taken up here: The prevention of radio interference by bonding, and the suppression of arcs.

1.8.1.1 BONDING

Any two points on the metallic structure of the airplane, whether electrically connected or not, may develop a potential difference at some frequency. At those frequencies for which the dimensions of the structural member are of the order of magnitude of a wave length, such potential differences are impossible to avoid in the presence of an electric or magnetic field. At the lower frequencies, the circuit concept may be used to show that the potential difference between two points of the structure is proportional to the impedance between the same points. Reducing the impedance will therefore reduce the potential difference at all frequencies at which this impedance may be considered as one lumped element.

Between two points which are in an electromagnetic field and which are electrically insulated from each other, there will exist a comparatively strong electric, but weak magnetic field, the latter being caused by displacement currents only, which are negligible at frequencies below about 100 megacycles. When the two points are "bonded", i.e., connected through a path of low impedance, a conduction current will exist with which is associated a comparatively weak electric, but strong magnetic field. The conduction current with its magnetic field is much less important as an interference generator than the electric field between insulated points for the following reasons: (1) When the two points are insulated from each other, even a small amount of charge accumulated at the points may cause a large potential difference. When the points are connected permanently, no charge can accumulate and the resulting steady state current is usually negligible; (2) If the points are permanently bonded, the impedance between them is much more likely to be constant than if the points are separated by a distance which may vary with any mechanical shock or other random motion of the structure. Thus, bonding will eliminate the generation

of interference caused by a varying impedance; (3) If the points are insulated from each other, the electric field between them may become large nov h to cause an arc or spark discharge of the accumulated charge. Arcs and sparks, as pointed out in Paragraph 1.3.2.4, are among the most serious sources of radio interference. This type of arcing is eliminated by proper bonding.

Bonding of the aircraft shell serves many purposes other than the elimination of interference, such as to provide a low impedance path for all electrical equipment that use the airplane structure for a return circuit, to minimize lightning damage, and to provide an effective antenna counterpoise. Therefore, good bonding will receive careful attention by the designer quite apart from any consideration of radio interference. It should be remembered, however, that a truly low impedance path is possible only so long as the dimensions of the bonded members are small as compared to a wave length of the interfering signal. At high frequencies, the members must be considered as transmission lines whose impedance may be inductive or capacitive and have any magnitude whatever, depending on their geometrical shape and the frequency. Bonding in itself, therefore, does not assure the existence of a true "ground plane" in the aircraft, i.e., the lack of an appreciable potential difference between any two points of the bonded members.

Bonding refers to the provision of a low impedance path not only between two points of the structure of the aircraft, but also between one point of the structure and a piece of equipment for which the structure serves as ground. Poor bonding of this kind is a very frequent cause of interference though it is not ordinarily considered as a source. Rather, poor bonding keeps other measures of suppression, such as the insertion of filters and proper "grounding", as discussed in Paragraph 1.8.2, from being effective. Examples of this will be found in Paragraphs 1.8.2.1 and 1.8.2.3.

1.8.1.2 SUPPRESSION OF ARCS

Arcs occurring during switching and other processes, which do not perform any useful function, must be prevented entirely both because they are serious sources of radio interference and because they produce a rapid deterioration of the contacts. Devices which help to extinguish the arc once it is established, such as the mechanical insertion of a dielectric between the contacts or the placement of a strong magnet close to the gap, are not suitable for radio-interference reduction because shortening the duration of the arc does not reduce its effectiveness as a source of high-frequency disturbances. The only suitable methods are those which prevent the formation of the arc at the outset.

By far the most effective method of preventing arcing across the contacts of a switch is the insertion, in parallel with the switch, of a capacitance in series with a resistance. The purpose of the capacitor is to prevent the voltage across the contacts from building up sufficiently to produce arcing; the resistance helps to damp out any oscillations that may occur and provides a means of dissipating the energy stored in the magnetic field of the current that was flowing before the switch was opened. To understand these actions, consider the circuit of Figure 1.8.1.2-A. In this diagram, E is a direct-current voltage source, R_1 is the load resistor which is connected to the source when the switch, S, is closed. Since every circuit contains inductance, L cannot be zero though it may be small. The quantity C_1 is the capacitance

between the contacts of the switch when open, and R_2 and C_2 are the elements of the external network added for the purpose of the arc suppression. In general, C_1 is not constant since it depends on the position of the contacts. But if R_2 is not too large and C_2 is much larger than C_1, then C_1 may be neglected.

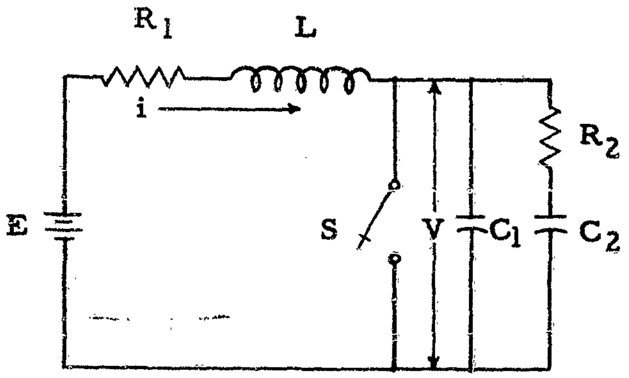


Fig. 1.8.1.2-A Switching Circuit with Arc-Suppressing Network

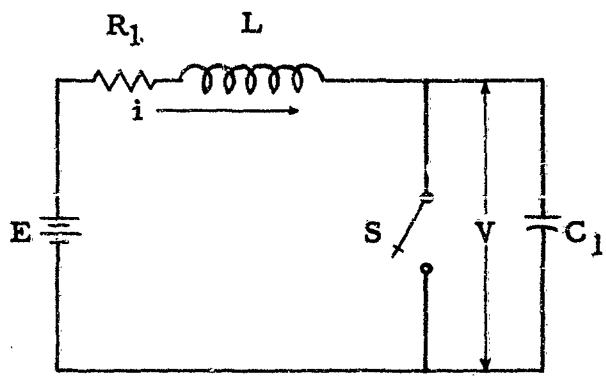


Fig. 1.8.1.2-B Simple Switching Circuit

It is assumed that the switch, S, is initially closed so that a steady current $I = E/R_1$ flows through it and there is no voltage across either capacitor nor any current through R_2. If then, at time $t = 0$, the switch is suddenly opened, the question is: What happens to the voltage, V, across the switch?

Before answering this complicated question, a simpler one will be considered: What happens to the voltage across the switch in the simpler circuit of Figure 1.8.1.2-B, in which the external arc-suppressing network has been omitted? The integro-differential equation for this circuit, for all times subsequent to the opening of the switch, is as follows:

$$R_1 i + L \frac{di}{dt} + \frac{1}{C_1} \int_0^t i \, dt = E \tag{26}$$

where i is the instantaneous current, and the initial condition is that $i = I$ at $t = 0$. The solution of this equation may be written in the following form:

$$i = \left(I \cos \omega t + \frac{E}{2\omega L} \sin \omega t \right) e^{-\frac{R_1}{2L} t} \tag{27}$$

where $\omega = \sqrt{(1/C_1 L) - (R_1/2L)^2}$

From this it follows that:

$$V = E + \left[\left(I\omega L - \frac{ER_1}{4\omega L} \right) \sin \omega t - E \cos \omega t \right] e^{-\frac{R_1}{2L} t} \tag{28}$$

Both i and V show damped sinusoidal oscillations provided that ω is real. To estimate the maximum voltage that can build up across the switch, assume that R_1 is

small so that there is little damping. Then the maximum value of V is approximately equal to $I\omega L = I\sqrt{L/C_1}$. Thus, if there were no capacitance across the contacts, the voltage would become infinite.

To return now to the circuit of Figure 1.8.1.2-A, let the capacitance C_1 be neglected since it is in parallel with C_2, assumed to be much larger than C_1. Then the circuit is similar to the simple one analyzed before with the capacitance C_2 substituted for C_1 and the total resistance $R_1 + R_2$ substituted for R_1. The major difference is that the voltage across the contacts is now the voltage across both R_2 and C_2 instead of being the voltage across the capacitance alone. The maxima of the voltages across R_2 and C_2 do not occur at the same time so that the maximum voltage across the switch is not equal to their sum. Nevertheless, in order to keep the contact voltage small, both these voltages must be small.

Proceeding on the same basis as before, to keep the voltage across C_2 small for a given I and L, this capacitance should be as large as possible. In order to keep the voltage across R_2 small, this resistance should be as small as possible since the maximum value of this voltage is simply IR_2, using the same approximations as before. It may be noted here that for a value of R_2 equal to $\sqrt{L/C_2}$ the maximum voltages across C_2 and R_2 become equal.

On the other hand, to increase the damping the total resistance of the circuit should be high. If weakly damped oscillations of large amplitudes are permitted to occur, they may produce as much radio interference as the arc that was suppressed. In fact, to prevent oscillations altogether, the total resistance should be sufficient for "critical damping", i.e., sufficient to reduce ω to zero or even make it imaginary. (A sine or cosine function with imaginary argument leads to an exponentially decaying function in this case.) The expression for ω shows that, for critical damping, the resistance should be equal to $2\sqrt{L/C_2}$, or just twice the value obtained before. If R_1 is very small, there could be no choice of R_2 that would satisfy both requirements, that of large damping as well as that of small contact voltage. Fortunately, in most practical cases, R_1 by itself is large enough to provide the damping, and a value of R_2 as little as 1/100 of the value required for critical damping may be a satisfactory choice.

The choice of C_2 is determined by the values of I and L, which must be known and can easily be measured, and by the values of the breakdown voltage between the contacts. The value of C_2 should be large enough to keep the maximum voltage, $I\sqrt{L/C_2}$, below that breakdown voltage by a wide margin of safety.

In determining the breakdown voltage between contacts, it must be remembered not only that tabulated values of dielectric strength (which is the ratio of breakdown voltage to the thickness of the dielectric between contacts) are very approximate and that the actual breakdown voltage is a function of many factors such as shape of the contact points, frequency and wave shape of the applied voltage, humidity, and temperature, but also that there is a definite and rapid decrease of the breakdown voltage with altitude due to the decrease of pressure and increase of ionization. The ratio of the breakdown voltage at any altitude to the breakdown voltage at sea level, other conditions being kept equal, is shown in Figure 1.8.1.2-C.

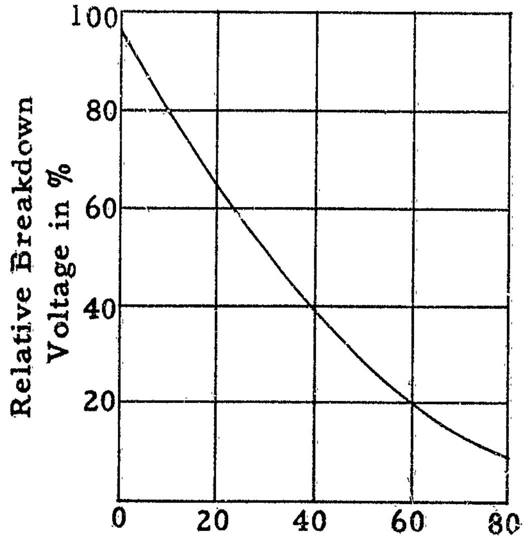


Fig. 1.8.1.2-C Effect of Altitude on Breakdown Voltage.

1.8.2 SUPPRESSION DURING TRANSMISSION

Once an interfering signal is generated, it must be prevented from reaching the receiver. Five methods are available for this: (1) location, (2) orientation, (3) grounding, (4) shielding, and (5) filtering.

The first two involve extremely simple principles, which can be stated in a few sentences. The other three require more detailed analysis and will be treated in later paragraphs.

Prevention of the transmission of interfering signals by proper location simply means that equipment likely to generate interference or lead wires likely to carry interfering currents should be mounted or placed as far away as possible from all receivers and all power, control, input, or output leads connected to any receiver. It also means that all equipment should be placed so as to make the maximum use of natural metallic shielding afforded by structural members such as the skin of the aircraft, bulkheads, and firewalls. Because of the high sensitivity of modern receivers, because of the possibility of resonance excitation (See Paragraph 1.6.4), and because of the possibility of direct conduction (see Paragraph 1.6.3), proper location cannot prevent all interfering signals from reaching and affecting the receiver, but it aids greatly in eliminating some and reducing most of the remaining interfering signals at the receiver. The proper location of equipment and wiring for minimum transmission of interfering signals is the responsibility of the systems design and lay-out engineer.

Prevention of the transmission of interfering signals by proper orientation of wiring means the utilization of the fact that the inductive coupling between two circuits can always be reduced to zero by proper orientation of one circuit with respect to the other. One example of this was given in Paragraph 1.6.1. Another example is illustrated in Figure 1.8.2. Here a comparison is made between the inductive coupling of two circuits that contain a section of parallel wires and that of two circuits having wires crossing perpendicularly. The amount of inductive coupling between the two circuits depends on the amount of magnetic flux of one circuit linking with the other. Figure 1.8.2 shows that the number of flux linkages is a maximum when the two wires run parallel, and zero when they are perpendicular.

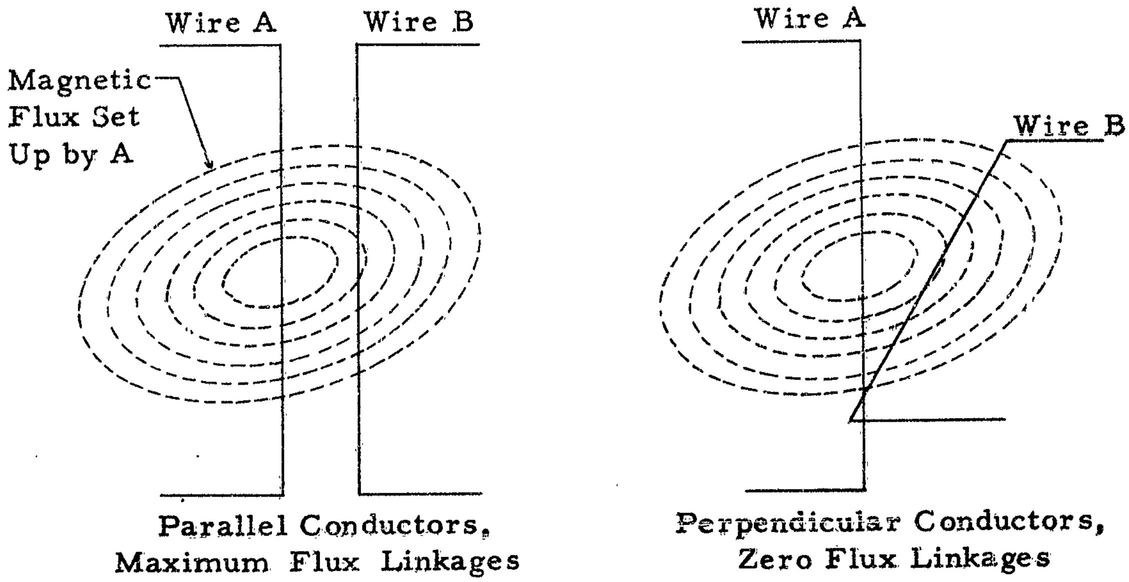


Fig. 1.8.2 Coupling of Long Straight Wires

Therefore, whenever conductors carrying interference currents and conductors sensitive to interference pickup must be close together, every effort must be made by the systems design and lay-out engineer to avoid paralleling them and to have them cross as nearly as possible at right angles.

1.8.2.1 GROUNDING

The word "grounding" has two different meanings in connection with aircraft electrical systems. Sometimes it is used synonymously with bonding to mean; connecting a point to the aircraft structure through a low impedance path. At other times it is used to mean; connecting a point electrically in such a way that it becomes equipotential with all other "grounded" points in the system. It is this second meaning which gives rise to much confusion, and many radio interference problems can be traced directly to a faulty interpretation of this concept.

It is a common procedure in aircraft wiring and circuit diagrams to use the ground symbol to indicate that a point should be connected electrically to the aircraft structure. This procedure is unambiguous for direct and low frequency power currents, but objectionable for radio frequency currents. As was pointed out and emphasized in Paragraph 1.8.1.1, bonding to the structure does not, in itself, assure the existence of a true ground plane at most of the frequencies encountered in interference problems. And it must be remembered that the impedance between two points supposedly at the same "ground" potential need not be very large to cause interference to be transmitted to the receiver.

Consider, as an example, the circuit of Figure 1.8.2.1. There are four different "grounds", one for the receiver, one to which the antenna is coupled capacitively, one for the motor, which is a source of interference, and one to which the by-pass condenser of the motor is connected. These four points are labeled 1, 2, 3, and 4, respectively. Assume that points 1 and 4 are at the same potential and also points 2 and 3, but let there be a small impedance between points 2 and 4 as indicated by the dashed connection. This impedance, Z, may be caused by poor bonding or by the fact that the structural member between these points has a small effective resistance or inductive reactance at the frequency of interest. Inspection of the

diagram shows that this impedance is common both to the interfering currents from the motor and the desired antenna currents in the receiver. Because of the high sensitivity of the antenna circuit of the receiver, even a very small interfering voltage developed across Z will cause an appreciable interfering signal at the output of the receiver. This example illustrates a very frequent cause of interference in aircraft.

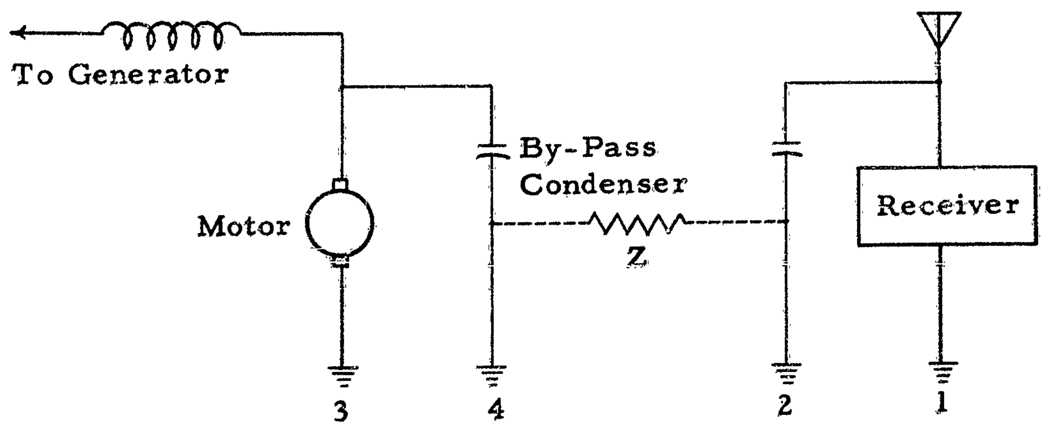


Fig. 1.8.2.1 Ambiguity of Ground Points

The designer of aircraft equipment and systems and the installation engineer must always be conscious of the possible ambiguity of ground points in the electrical system. He must keep clearly in mind the fact that proper "grounding" is not entirely a matter of good bonding, and that it may be impossible to have two points at the same potential over an appreciable range of frequencies unless the points are physically close together.

1.8.2.2 SHIELDING

A shield is a partition between two regions of space such that the electric and magnetic fields of interest are attenuated in passing from one of these regions to the other. All practical shields are made of metals of high conductivity.

The shielding action of metallic sheets may be explained in two ways depending on whether the field or circuit concept is used. According to the field concept, the shielding action of a metallic sheet is two-fold: An electromagnetic wave striking a metallic surface is partially reflected, and the transmitted part is attenuated in its passage through the sheet. According to the circuit concept, the currents flowing in circuits on one side of the sheet induce currents in the sheet. The induced currents produce fields on the other side which just cancel the fields due to the original currents. Mathematical analysis on either basis is very difficult, and is practical only for cases of simple geometry such as shields in the shape of infinite plane sheets or infinitely long circular cylinders. The second example finds application in the shielding of cables, but most shielding problems in aircraft are not amenable to simple mathematical analysis. A compilation of the most important analytical expressions for the effectiveness of shielding in certain simple cases together with their derivations is given in Appendix XVI.

One of the most significant results is that a plane electromagnetic wave is attenuated very rapidly in a metallic medium after entering it through a plane boundary

surface. The fields decrease exponentially according to the law

$$F = F_o \, e^{-1.238 \sqrt{\frac{\mu f}{\rho}} S} \qquad (29)$$

where F is the electric or magnetic field intensity at a distance S inches away from the surface, F_o the same field intensity at the surface in the same units as F, ρ the resistivity of the material in ohm-circular-mils per foot, μ the relative permeability (μ =1 for all non-magnetic materials), and f the frequency in cycles per second. This variation is shown in Figure 1.8.2.2. The above equation is exact only when the metal extends to infinity in the direction of increasing S, but in case of a shield of finite thickness the equation is a good approximation when the field at the far end is sufficiently small. In practice, even a very thin metallic sheet will allow only a small fraction of the entering energy to pass through it. In addition, the reflection coefficients of most metallic surfaces to plane waves are large so that only a small fraction of the energy striking the surface will enter the metal. If a shield must support itself mechanically, the thickness required by mechanical considerations is usually in excess of that required for effective suppression of a plane wave except at low and very low frequencies. However, this does not apply to metallic coatings applied to mechanical supports of other materials. These coatings are often so thin that no effective shielding action is obtained. Of course, even a comparatively thick shield might not offer sufficient attenuation if the source is very strong and the receiver very sensitive.

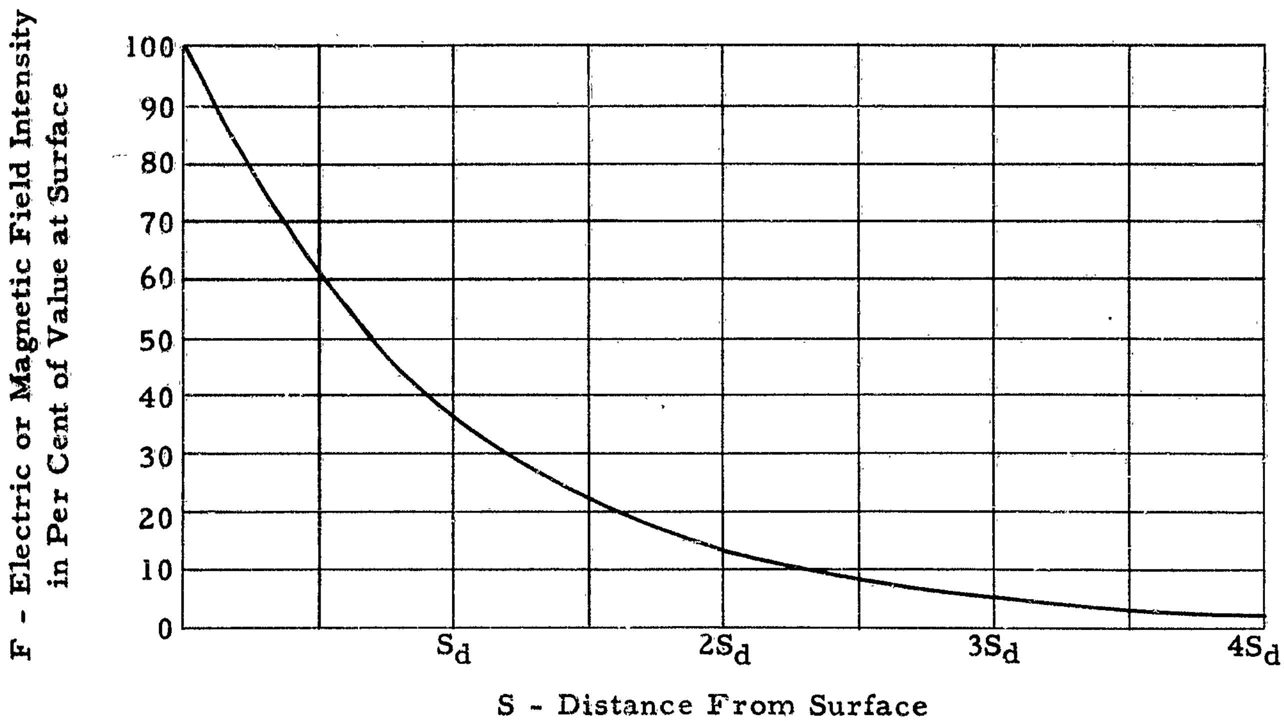


Fig. 1.8.2.2 - Variation of Electric or Magnetic Field Intensity Near the Surface of a Plane Conductor

It is often convenient to define the "depth of penetration" of a plane wave for a plane sheet. This is that value of S, called S_d, which makes the exponent in Equation

(29) equal to -1. Thus

$$S_d = \frac{1}{1.238\sqrt{\frac{\mu f}{\rho}}} = 0.8079\sqrt{\frac{\rho}{\mu f}} \tag{30}$$

Thus, the depth of penetration is the thickness through which the wave must travel to be attenuated to 1/e or about 37% of its original amplitude.

The depth of penetration of an electromagnetic wave is a concept closely related to that of the "skin effect". Just as the wave, at high frequencies, is very rapidly attenuated as it progresses into the metal, so the current decreases with the distance from the surface. This crowding of the current near the surface of a conductor is called the skin effect, and it gives rise to the increase of the effective resistance of a conductor with frequency. The shielding action of a metallic sheet may also be explained in terms of the skin effect. If the shield is thick enough so that the currents induced in one side of the sheet are practically reduced to zero on the other side, then no electromagnetic field can exist on the other side (provided that there is no source on that side) and the shielding action is complete. Since the current decreases exponentially, it actually never reaches the value of zero no matter how thick the shield is. This shows clearly that there cannot be a perfect shield theoretically. However, at higher frequencies the decay is so rapid that the shield need not be very thick to make the fields on the other side practically undetectable by the most sensitive receivers available.

Equation (29) shows that the effectiveness of shielding, as far as absorption within the metal is concerned, is directly proportional to the thickness of the shield, to the square root of its conductivity, and to the square root of its magnetic permeability. It also shows that the shielding effectiveness increases as the square root of the frequency so that the thickness, conductivity, and permeability of the shielding material should be chosen for the lowest frequency. However, when magnetic materials are used, it must be remembered that the permeability of most magnetic substances decreases with frequency. Therefore, an increase of shielding effectiveness with frequency is not always realized. It should also be noted that, as far as openings and joints in shields are concerned, their "leakiness" depends on their dimensions measured in wave lengths, hence their presence makes the shielding effectiveness decrease with frequency. More specific practical rules will be given in Paragraph 3. 1. 2.

The discussion above, together with Equations (29) and (30), is based on plane waves. It is shown in Appendix XVI that within metals electromagnetic fields behave like plane waves under a variety of different conditions. For example, the remarks about the skin effect and Equation (30) are applicable to most metallic surfaces as a good approximation even for cylindrical waves unless the surface is curved with a radius of curvature much smaller than a wave length. On the other hand, the fields encountered in radio interference problems are often not associated with plane waves. The actual fields in the vicinity of sources of interference may be so complicated that the application of the results obtained on the basis of plane waves may lead to serious errors.

Practical shields are usually designed so as to enclose completely either the source in order to keep the interference in, or the receiver in order to keep the interference out. In either case, assuming that the shielding material has sufficient thickness, conductivity, and permeability, the effectiveness of the shield depends on its completely enclosing the source. The inside (or outside) of the shield must form a continuous surface of lower impedance than any other possible current path leading to the outside (or inside) of the shield. This means that openings must be avoided, and any joints must be carefully designed so as to make sure that <u>good electrical contact is made along a continuous line.</u> In ignition systems, which depend entirely on shielding for interference-free operation, practically all interference troubles have been traced directly to faulty joint design. When openings are necessary, as for ventilating purposes, they must be specially designed for minimum transmission of interfering signals. They must interrupt the flow of induced currents in the shield as little as possible, and they must strongly attenuate any radiation through them. Protruding sleeves around the openings, acting as wave guides below their cut-off frequencies, have been found very helpful.

When long cables are shielded, the outside of the shield is another possible path for interfering currents from sources that may have no connection with the currents within the shield. To minimize this possibility as well as the effect of any faulty shield or shielding joint design, all cable shields must be properly grounded at least at each end, and for very long cables also at intermediate points. It might be argued that "floating" shields, or shields grounded at only one point, are also effective in preventing undesired currents through them. But it has been found in practice that grounding as suggested above is more effective despite the limitations pointed out in Paragraph 1.8.2.1.

Certain types of special purpose shields are sometimes used to reduce one type of coupling without affecting another. The most important example of this is the "Faraday shield", which is used to prevent capacitive coupling between two coils without affecting the inductive coupling. A Faraday shield is a set of grounded metallic prongs, arranged somewhat like the teeth of a comb, placed between two coils. Since the prongs are not connected at one end, no induced currents can flow through them and the magnetic coupling is not affected. But the prongs are so close together that their plane is essentially equipotential. Thus, no electrostatic coupling can exist between the two coils. Another example of this is the electrostatic shield around a loop antenna, used to make the antenna responsive only to magnetic fields. This device is useful when the most important interfering signals have a much larger ratio of electric to magnetic field intensity than the desired signals.

1.8.2.3 FILTERING

No matter how well a source is shielded, some electrical connections must be made to it that will break the shield because energy must be either supplied to it or carried away from it or both. In addition, control cables may have to be connected to it. These power and control leads will conduct interfering currents away from the source despite all efforts toward perfect shielding. The conduction of interfering currents may be impeded by the use of filters or other special networks. The use of special filter-like networks with dissipative elements will be discussed in Paragraph 3.1.1.7.

A filter is a four-terminal network designed to freely transmit currents or voltages of certain frequencies while attenuating all others. To do this, use is made mainly of reactive elements, i.e., inductances and capacitances. The presence of dissipative elements, that is, elements with effective resistance or pure resistances, prevents the free transmission of desired currents or voltages and is therefore usually avoided in filters.

Filters are classified according to the band of frequencies to be transmitted and attenuated as low-pass, high-pass, band-pass, and band-elimination filters. Typical examples of the simplest basic structures for each of these types, together with typical attenuation curves for each, are shown in Figure 1.8.2.3-A. The frequencies that separate transmission and attenuation regions are called the cut-off frequencies, f_c, of the filter as noted in the diagrams.

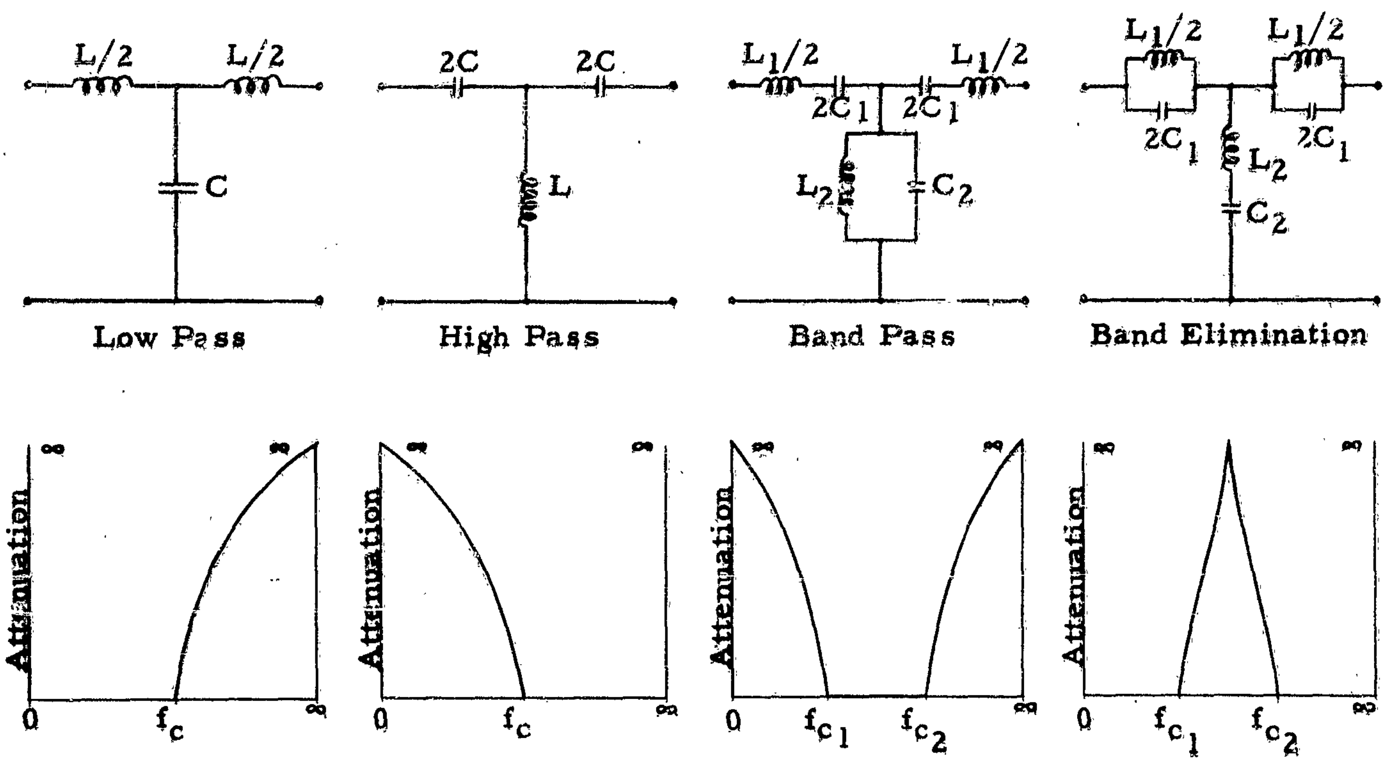


Fig. 1.8.2.3-A Typical Filter Sections Showing Attenuation As a Function of Frequency

The elements of the filter must be so chosen that the impedances looking into and out of the filter remain approximately the same as those of the transmission line into which it is to be inserted at the frequencies it is desired to transmit. This is necessary in order to insure that the filter does not impair the normal functioning of the equipment at both ends of the transmission line. In other words, the load impedance as seen by the generator should not be changed by the insertion of the filter so that the generator still delivers the current for which it was designed at the voltage for which it was designed. And the load should still be fed by a network of the same open-circuit voltage and the same internal impedance so that it operates exactly in the way intended by the designer. On the other hand, at the frequencies the filter is designed to attenuate, the impedances as seen by the generator and the load are different from what they were before the insertion of the filter - usually either very high or very low.

In most practical aircraft interference-suppression problems, the main consideration is to pass freely a power or control current comprising a single frequency or at most a very narrow band of frequencies, while at the same time greatly attenuating all other (interference) frequencies. In many cases this can be accomplished by designing a simple low-pass filter in accordance with the data given for element values in Appendix VII. The details to be considered in constructing the capacitors and inductors are given in Paragraphs 3.1.1.4 and 3.1.1.6. The designer of interference suppression filters usually will have a wide choice of parameters at his disposal depending upon the ratio of the lowest frequency to be suppressed and the frequency to be transmitted, and also the attenuation requirement to be met. For example, if the interfering frequencies to be suppressed are above 100 kc while the frequency to be passed is 400 cps, the cut-off frequency of a low-pass type filter could be chosen anywhere in the approximate range from 500 cps to 90,000 cps depending on the number of decibels suppression required at 100,000 cps and above. The values of inductance and capacitance can, therefore, vary over a wide range but the ratio of their values is fixed by the impedance desired at 400 cps, because it is this ratio which determines the impedance looking into the filter and, therefore, the attenuation loss suffered by the desired 400 cps signal.

The design of filters is an art as well as a science since so much depends on the judgment and techniques used by the filter design engineer. However, the formulas, curves, and tables given in Appendix VII will assist the design engineer in the solution of interference suppression problems on a scientific basis and thus avoid the pitfalls which are almost certain to be encountered in choosing values of elements on a trial and error basis.

In some cases the designer may wish to choose a band-pass filter when it is desired to transmit a band of frequencies without attenuation and distortion while at the same time highly attenuating all frequencies above and below this band. For example, it may be desired to transmit voice frequencies along an intercommunication system in aircraft, or the input circuit to a receiver may require a broad band filter. Also, circuits within a receiver itself, such as audio output or intermediate frequency circuits, often require faithful transmission over a fairly wide band of frequencies while suppressing all other frequencies. In these cases the designer first of all determines the approximate impedance level of the line into which the filter is to be inserted, the upper and lower cut-off frequencies, and the attenuation required for all other frequencies. By use of the formulas and charts given in this book, the correct values of elements can then be readily computed for maximum performance.

Unless the ratio of the upper to the lower cut-off frequencies is less than 2 to 1, it will usually be found advantageous to employ high-pass and low-pass filter sections in series. This advantage will be apparent if the size of the inductors and capacitors are computed both for band-pass and for low or high-pass sections.

It should be pointed out that there may be other good reasons for choosing a band-pass type filter, as, for example, to make use of impedance transformations to increase or decrease the current or voltage in the line (to avoid large magnetic or electric fields), or to obtain more easily realizable values of elements. The techniques used by filter design engineers to accomplish these results are beyond the scope of this volume and, therefore, cannot be included in this edition of the book.

It must be pointed out that there are many cases in which the impedance of the filter at the frequency to be transmitted is of secondary importance. If the current to be passed is direct current, i.e., zero frequency, then a low-pass filter does not change the impedance level at all looking either way, except for the effects of possible resistance in the coils. Also, neither a series inductance, whose impedance is zero at zero frequency, nor a shunt capacitor, whose impedance is infinite at zero frequency, in any way affects any direct currents or voltages in a system. It can be shown that this statement is substantially true also for low frequency alternating currents, provided only that the frequency to be transmitted is sufficiently removed from the cut-off frequency of the low-pass filter. To show this, the effect of the insertion of a symmetrical filter into a circuit is considered. Let a generator of voltage E having an internal impedance Z_s be connected directly to a load of impedance Z_R. Then the current is $E/(Z_s + Z_R)$ at all frequencies. After inserting a filter to suppress the radio interference frequencies, it is essential that the current at the desired frequency or frequencies be the same as before, but currents at all other frequencies be much less. If this can be accomplished, then the functional operation of the generator and the load is the same as before the insertion of the filter. Now let a filter, with image impedances Z_{I_1} and Z_{I_2} and with image transfer constant θ, be inserted as shown in Figure 1.8.2.3-B. Since the filter is assumed to be symmetrical, $Z_{I_1} = Z_{I_2} = Z_I$.

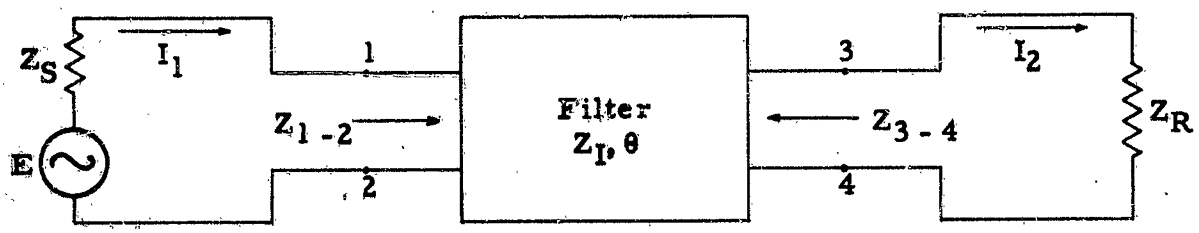


Fig. 1.8.2.3-B Insertion of Filter Between Generator and Load

The input current, I_1, is given by Equation (10) in Paragraph 1.4. From this equation, the impedance, Z_{1-2}, looking into the filter terminated in Z_R, may be obtained and is given by:

$$Z_{1-2} = Z_I \left[\frac{1 + F_r' \, e^{-2\theta}}{1 - F_r' \, e^{-2\theta}} \right] \tag{31}$$

where $F_r' = (Z_R - Z_I) / (Z_R + Z_I)$. Since it is required that the impedance as seen by the generator should still be Z_R at the desired frequency, the important relationship is that between Z_{1-2} and Z_R. The expression for the input impedance may be changed to the following form:

$$Z_{1-2} = Z_R \left[\frac{(1 + e^{-2\theta}) + \frac{Z_I}{Z_R}(1 - e^{-2\theta})}{(1 + e^{-2\theta}) + \frac{Z_R}{Z_I}(1 - e^{-2\theta})} \right] \tag{32}$$

To obtain a clearer picture of the effect on the input impedance, assume that θ is small as compared to unity, so that $e^{-2\theta} \simeq 1 - 2\theta$. Then,

$$Z_{1-2} \simeq Z_R \frac{1 + \left(\frac{Z_I}{Z_R} - 1\right)\theta}{1 + \left(\frac{Z_R}{Z_I} - 1\right)\theta} \simeq Z_R \left[1 + \theta\left(\frac{Z_I}{Z_R} - \frac{Z_R}{Z_I}\right)\right] \tag{33}$$

It is seen that $Z_{1-2} = Z_R$, i.e., the impedance as seen by the generator remains unchanged, when either $\theta = 0$ or $Z_R = Z_I$.

In a low-pass filter, both θ and Z_I vary with frequency. In the pass band, θ is imaginary, $\theta = j\beta$. The phase shift, β, is given by:

$$\beta = \tan^{-1}\left[2\frac{f}{f_c} \frac{1 - \left(\frac{f}{f_c}\right)^2}{2\left(\frac{f}{f_c}\right)^2 - 1}\right] \tag{34}$$

where f is the frequency and f_c is the cut-off frequency of the low-pass filter. If f/f_c is small, β is also small, and one may write, approximately,

$$\beta \simeq -2\frac{f}{f_c} \tag{35}$$

The image impedance of a constant-k low-pass filter varies with frequency as follows:

$$Z_I = R\sqrt{1 - \left(\frac{f}{f_c}\right)^2} \simeq R \tag{36}$$

where R is the design resistance of the filter. Combining these equations, one obtains:

$$Z_{1-2} = Z_R\left[1 + 2\frac{f}{f_c}\left(\frac{R}{Z_R} - \frac{Z_R}{R}\right)\right] \tag{37}$$

This shows that, in order to keep Z_{1-2} substantially equal to Z_R, the term

$$2\frac{f}{f_c}\left(\frac{R}{Z_R} - \frac{Z_R}{R}\right) \tag{38}$$

must be kept small. To do this, neither f/f_c nor R/Z_R (or Z_R/R) can be large.

In Figure 1.8.2.3-C, the quantity Z_{1-2}/Z_R is plotted as a function of the mismatch-ratio, R/Z_R, for different values of f/f_c. For example, it is seen that, when

$f = 400$ cps and $f_c = 40$ kc, so that $f/f_c = 0.01$, the ratio R/Z_R may be as small as minus one and as large as ten without changing the impedance level more than 20 percent. If a change of no more than 10 percent is desired, either the cut-off frequency must be raised or the ratio R/Z_R must be made closer to unity.

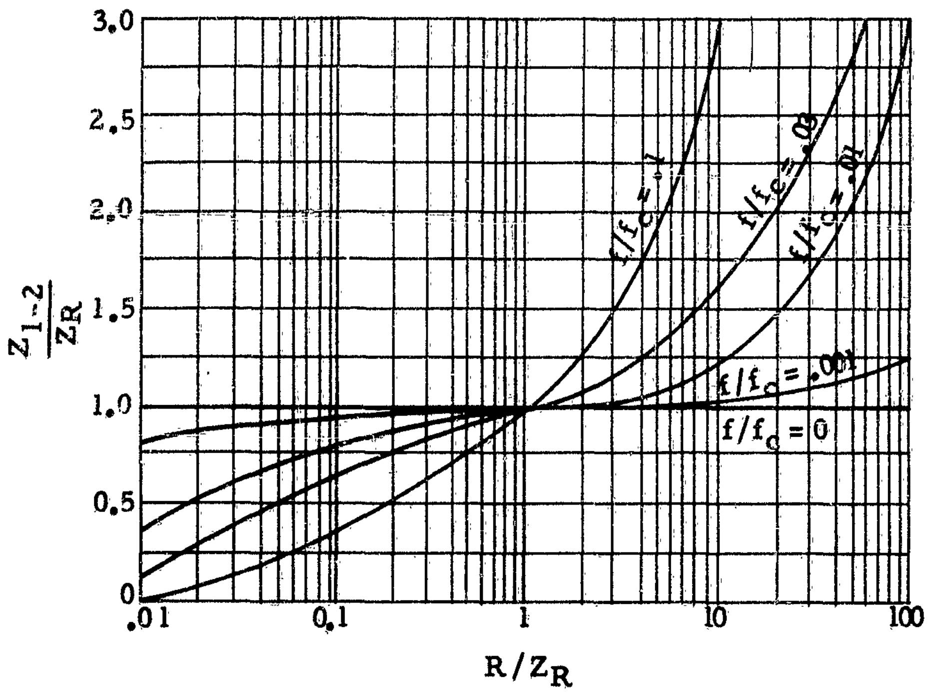


Fig. 1.8.2.3-C The Effect of Mismatch on Input Impedance of Low-Pass Filter

The equations derived hold equally well for both sides of the filter. From the relationships given, it follows that, if the ratio of Z_s to Z_R is very high or very low, it is most desirable to choose R near the geometric mean, $\sqrt{Z_s Z_R}$, in order to keep both R/Z_R and R/Z_s as close to unity as possible.

When the only desired mode of transmission is direct current, that is, when the frequency of the desired current is zero, a simple shunt capacitor or series inductance or combination of the two may be effective in suppressing interference frequencies from zero to infinity. Such a network is often called a "brute-force filter" because it does not satisfy the definition of a filter as stated above. Since these networks are often used to suppress interference in circuits where only direct currents are to be passed, they are usually designed without regard to specific frequency or impedance requirements because the frequency of the pass band is zero and the impedance is important only for the reasons discussed below. The design of these networks is treated in Paragraph 3.1.1.1.

Typical arrangements of elements for "brute-force" filter networks are shown in Figure 1.8.2.3-D.

The input impedance of any filter, including that of the so-called "brute-force" type, as seen by the source, is normally either very high or very low over most of the attenuation (interference) band as compared to that required for a good match in the transmission band. This fact has an important consequence for the behavior

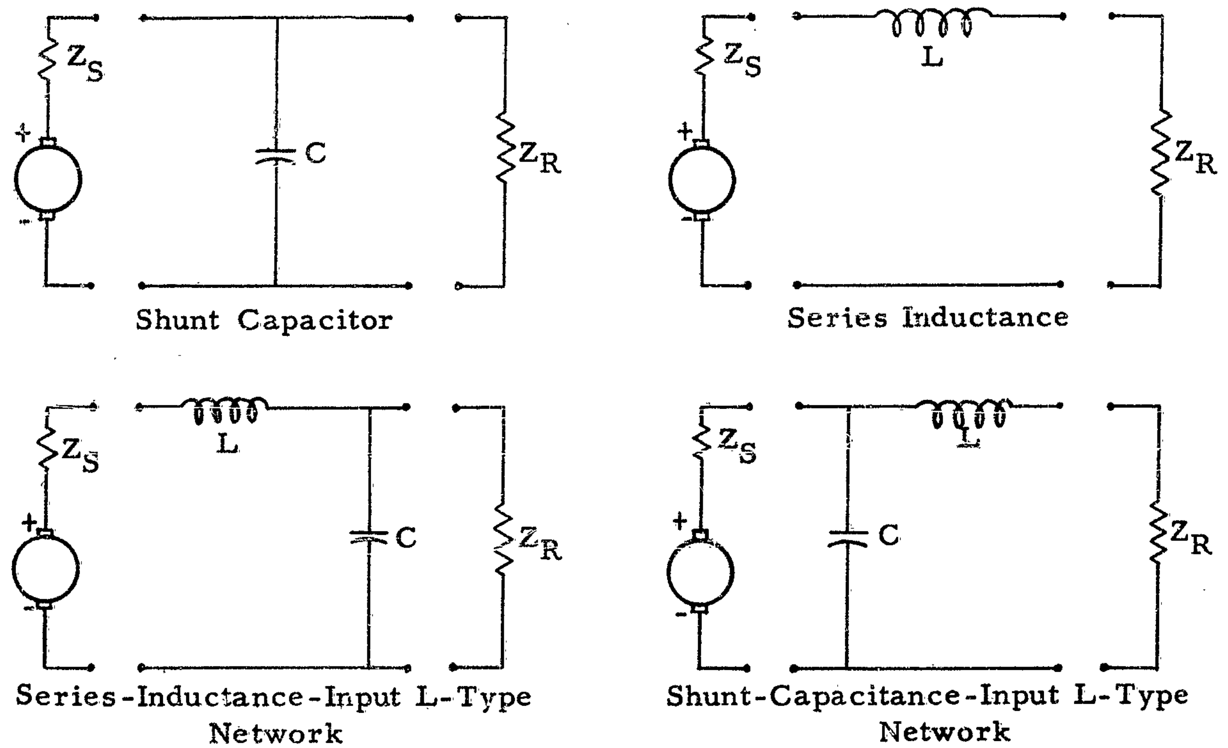


Fig. 1.8.2.3-D "Brute-Force Filters"

of the circuit before the point at which the network is applied. If the source of interfering signals has a low internal impedance at those frequencies, a low-impedance-input network increases the amplitude of the interfering currents between it and the source. If there is an appreciable amount of coupling between these interfering currents and a receiver through, say, magnetic induction, the suppressing network may increase the interference rather than decrease it. If, on the other hand, the source has a high internal impedance, a high-impedance-input network increases the voltage across its input. As before, this voltage may affect a receiver through, say, capacitive coupling, so as to increase rather than decrease the interference. In practice, the source may at times have a low internal impedance and at other times have a high internal impedance, depending on the frequency or mode of operation. Therefore, great care must be used by the designer in the choice of suppressing networks to insure a reduction of all types of interference at the receiver and to avoid the pitfalls of reducing the effect of conducted interference at the expense of interference coupled to the receiver in other ways.

The location of the filter or other suppressing network is another item that must receive careful consideration. Possible choices are shown in Figure 1.8.2.3-E. It is clear that one filter at each of the possible input paths of the receiver will take the place and do the work of many filters at the outputs of all the interference sources. On the other hand, the effect, mentioned before, of increasing interfering currents or voltages before the point of application of the filter, obviously increases as the filter is moved further from the source and is minimized by placing the filter directly at the source. Care must also be taken that filter coils and leads are not themselves placed in interfering fields. For example, by bundling the output leads of a filter together with the input leads, or by placing the coil of a filter close to a

coil carrying interfering currents, the effect of the filter may be completely nullified. To avoid this last possibility, it may be necessary to shield the filter coil or perhaps use special core designs. All these factors must be considered in arriving at a decision as to the best location of the filters.

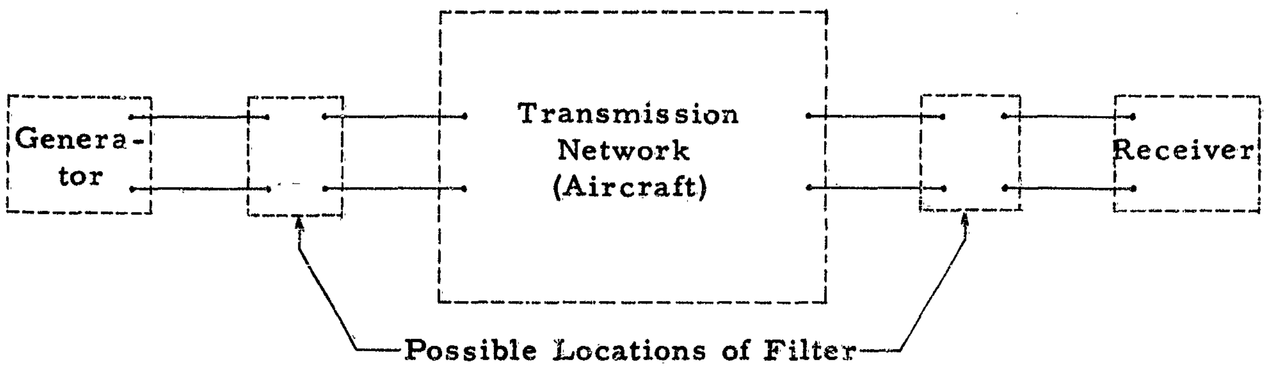


Fig. 1.8.2.3-E Location of Filters

Detailed filter design will not be considered since it has been treated adequately in several text books, as noted in the list of Selected References, Appendix III. The most important design formulas are listed in Appendix VII for easy reference. The limitations of these formulas, however, should be clearly recognized. They lie chiefly in four facts: (1) the resistance of inductance; (2) the distributed capacitance of inductance coils; (3) the distributed inductance of capacitor leads; and (4) stray coupling between coils or leads.

The first of these limitations, the resistance of inductance coils, is especially troublesome for direct currents since it may cause an excessive voltage drop across the coil and a corresponding reduction of useful output. It is also of great importance in the design of "wave traps" where a very narrow band of frequencies is to be attenuated. It is also of importance in obtaining sharp cut-offs in filters; but, as stated elsewhere, most filters used for interference reduction do not require sharp cut-off characteristics. The resistance of a coil is usually measured in terms of a figure of merit, called the "Q" of the coil and defined as the ratio of the inductive reactance to the resistance:

$$Q = \frac{\omega L}{R} \tag{39}$$

where ω is the angular frequency in radians per second, L the inductance in henries, and R the effective resistance in ohms. It is seen that "Q" depends on the frequency, but at radio frequencies the resistance is a function of frequency also. Usually, over a limited range of radio frequencies, it is found that the "Q" of a coil is more nearly constant than its resistance. If the coil is used in combination with a capacitance, either in series or in parallel, then the "Q" specified usually refers to the resonant or anti-resonant frequency of the combination.

The second and third limitations noted above apply especially at higher frequencies. The distributed capacitance of an inductance coil may cause it to have a

negative reactance at frequencies as low as 10 kc, but this frequency may be pushed up considerably by proper design. The lead inductance of a capacitor may cause its reactance to become inductive, and for capacitors with external leads this usually happens between 100 kc and 10 mc, approximately. In the "feed-through" capacitors (see Paragraph 3.1.1.5), which have no external leads in series with the capacitance, this frequency may be pushed up to 1000 mc and higher. Obviously, filters will lose their effectiveness when their elements undergo such radical changes.

The fourth limitation, stray coupling, is troublesome at all frequencies. It may be inductive or capacitive. It occurs most frequently between the input and output leads of the filter, between the series arms of a T-network, and between the shunt arms of a pi-network. Specific methods of minimizing this limiting factor are outlined in Section 3.

If filters are to be constructed for frequencies that are too high for lumped elements, sections of transmission lines must be used instead. Because of the difficulty in constructing filters that are effective over very wide ranges of frequencies, it may be necessary at times to use two or more filters in series to cover the entire frequency band in several steps.

In designing filters for the suppression of radio interference, especially those for power and control currents, three aspects need special consideration. The first is that there is usually only one frequency to be transmitted and an extremely wide band of frequencies to be attenuated. The second is that the impedances of the terminal equipment are usually not controlled by the filter designer. They are not usually constant over a wide frequency range; and, moreover, they are usually unknown at all but the one desired frequency. The third is that filters, being constructed mainly of reactive elements, cannot dissipate the energy associated with the generation of radio interference.

In many cases of interference suppression in aircraft, the designer will be concerned only with an extremely narrow transmission band and an extremely wide attenuation band. Therefore, the sharpness of the filter cut-off and the behavior in the comparatively wide region between the frequency to be transmitted (direct current or 400 cps) and the lowest frequency to be suppressed (about 150 kc) are relatively unimportant. The only important requirements are minimum attenuation for one frequency and very large attenuation for all interfering frequencies, from 150 kc to 1000 mc, in the most general case.

The fact that the terminating impedances are not constant with frequency and are usually not even known quantities leads to considerable complications in filter design. Most filter design considerations, as found in standard text books, are based on constant-resistance terminations. In this case, no distinction need be made between transmission of power on the one hand, and that of currents or voltages on the other because, for a resistive termination, there exist the simple relationships $P = i^2R = E^2/R$, so that, when the power is zero, both current and voltage must be zero also. This is not true when the terminating impedances can be reactive and can even assume values of zero or infinity, i.e., become resonant or anti-resonant, at certain frequencies. Hence, it is possible for a filter to have an appreciable current or voltage output despite a high attenuation constant at a particular frequency.

As was pointed out in Paragraph 1.4, in radio-interference problems, it is not always the power output that is important, but more often the voltage or current output, even with very little power. Obviously, this aspect requires the special attention of the designer.

The fact that filters made of reactive elements cannot dissipate any energy leads to a considerable increase in the energy that must be dissipated in the network before the filter. This is not harmful if the filter is placed at the output of an interference source which is well shielded and has no other possible outlets for the interfering energy. In this case, the energy is completely "bottled up" - the filter acting as an effective stopper - and is dissipated inside of the shield, where it can do no harm. When the shielding is not complete, the reflection of the energy back into the source might be harmful and lead to increased leakage; but, it must be remembered that one function of a properly designed filter is to prevent the generation of real power at the frequencies to be attenuated, by offering a reactive impedance at these frequencies. If this function is properly performed, there is no need for increased dissipation of real energy. There exists apparent, or reactive, power, which fluctuates back and forth between the filter and the source and need not be dissipated.

Finally the extreme importance of good grounding of filters must be emphasized. A filter can be made completely ineffective by improper grounding techniques. Figure 1.8.2.3-F shows how a high impedance bond to ground offers the interfering currents an effective by-pass path, thus vitiating the purpose of the filter.

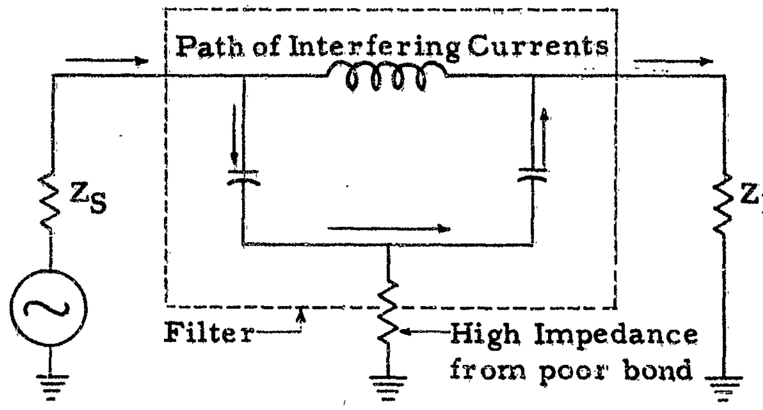


Fig. 1.8.2.3-F Effect of Poor Grounding on Filter Effectiveness

1.8.3 SUPPRESSION IN RECEIVERS

In spite of all efforts to suppress the interfering signal at the source and to prevent its transmission, there will always be some such signals that reach the receiver. In the design of receivers which will not be susceptible to interference, the designer must strive (1) to prevent the interfering signals from entering the receiver and affecting any sensitive circuits, and (2) to minimize the effects of the interfering signals in case they have gained entrance and have affected a sensitive circuit. The first is a problem of transmission. The second is a matter of utilizing, in one way or another, any differences that may exist between the desired and the interfering signal.

1.8.3.1 DESIGN FOR MINIMUM TRANSMISSION

The problems encountered in trying to prevent an interfering signal from entering the receiver have been discussed previously in Paragraph 1.8.2. They are problems of shielding the entire receiver including all antenna lead-ins, filtering all power and control leads leading into the receiver, and designing the receiver in such a way that interfering signals entering through the output leads cannot affect any sensitive portion of the receiver. This last problem is a difficult one since filtering in the output leads is not usually practical, and once an interfering signal has entered the receiver, the problems of preventing its transmission to a sensitive circuit are greatly magnified. Extensive shielding and filtering within the receiver may, at times, be necessary.

Attention must also be paid to the location of switching devices related to receiver operation. Often an antenna relay is found within the receiver case, thus affording the interfering signals from the transmitter an easy entry. The number of leads entering a receiver must be kept at the absolute minimum, and any device, not an integral part of the receiver, should not be placed in the receiver case.

1.8.3.2 BANDWIDTH CONSIDERATIONS

One difference between the desired and interfering signals is their frequency content. The desired signal contains only frequencies within a well-defined region of the frequency spectrum; for example, the carrier and two side-bands for double side-band transmission in amplitude modulation. On the other hand, the interfering signal is spread fairly evenly over a very large portion of the frequency spectrum. Of these frequencies, the receiver accepts only those that fall within its band of acceptance. Hence, a receiver should be designed so as to reduce the width of its acceptance band to the minimum required for the reception of the desired signal.

The last sentence must be interpreted in the light of the results of a statistical analysis of random noise. Such analysis shows that the larger the bandwidth of the receiver, the greater, potentially, could be the improvement of the signal-to-interference ratio provided the bandwidth is fully utilized for the improvement of this ratio. Hence, "minimum" means the smallest bandwidth required for making the fullest use of the potentialities of the system to improve the signal-to-interference ratio. Thus, for example, a receiver using frequency modulation will provide a greater signal-to-interference ratio at its output than one using amplitude modulation for the same signal-to-interference ratio at the input, even though it has by far the larger bandwidth. The reason for this is that the frequency-modulated receiver makes more efficient use of its bandwidth. This has been discussed in detail in Paragraph 1.7.4.

On the other hand, for limiters to be effective (see Paragraph 1.8.3.3.1), interfering pulses should undergo as little pulse-lengthening as possible. Hence, the bandwidth should be as large as possible. Thus, the bandwidth requirements are contradictory, and the actual design must be a compromise. The designer must decide, in each case, whether it is better to allow more interference to enter the receiver and suppress it later by effective limiting action, or to allow as little interference as possible to enter, with the knowledge that any attempt to limit later will be quite ineffective.

1.8.3.3 SPECIAL CIRCUITS

A large number of special circuits have been developed for the purpose of reducing radio interference in receivers. Only those circuits are treated in this paragraph which attempt to reduce the interference after it has entered the antenna or any other point along the normal route of the desired signal. In other words, the circuits to be discussed act after the interfering and desired signals have mixed, and therefore they must separate the two on the basis of some intrinsic difference between them. This difference may lie in the wave form, in the frequency distribution, or in some definite phase relations that apply to one, but not to the other.

The circuits to be discussed fall into one of five groups: (1) limiters, (2) wave traps, (3) blanking circuits, (4) phase-cancelling circuits, and (5) audio filters. The first and third are applicable when the interfering signal consists of large-amplitude pulses whose duration is very short as compared to their period. The second is applicable if the interfering signal contains only one, or at most only a very narrow band of radio frequencies. The fourth is effective for interfering signals whose character and path of entry are precisely known. Finally, the fifth is effective when the interfering signal contains only a small number of fixed audio frequency components. Of the five, only the first three are extensively used in receivers at this time.

1.8.3.3.1 LIMITERS

The action of an amplitude limiter is based on the fact that, in amplitude modulation, the amplitude of the desired signal varies from zero to, at most, twice the carrier amplitude, reaching that amplitude only during the rare modulation peaks. If an interfering signal contains pulses of short duration whose amplitude rises above that value, its effect will be greatly reduced if all amplitudes are limited to that same value. This limiting action does not affect the desired signal. Noise limiters of this type are quite effective and are now a standard part in many receivers.

Limiting action in these circuits is usually provided either by the switching action of a diode (in so-called "diode limiters") or by saturation of some tube. A diode limiter may be either the series or shunt type. In the series type, the diode is in series with the normal plate current flow so that it is conducting as long as the signal does not exceed twice the carrier level, but becomes non-conducting for a short-time interval when the signal exceeds this value. In the shunt type, the diode is in parallel with the normal plate current flow so that it is normally non-conducting, but becomes conducting for excessive signal amplitudes. In either case, the signal becomes greatly attenuated during the time that the limiting action of the diode takes place. Clearly, this time must be small enough so that the desired signal remains unaffected. Making use of the saturation effect in a vacuum tube has the advantage that no special tube is needed to provide noise limiting action, but that an existing tube may be utilized for this purpose. Practically, however, it has been found that this method is less effective and introduces greater distortion in the desired signal than a separate limiter stage.

Because of the pulse-lengthening action of the various stages of the receiver

(see Paragraph 1.7.2), the most desirable location of the noise limiter circuit is before the radio frequency amplifier, or at any rate, as close to it as possible. Diode limiters, however, need both direct current pulses and considerable voltage amplitudes for their operation, which normally are not available in the early stages of the receiver. The generally accepted location of diode limiters is therefore the second detector stage, where the necessary voltages and currents are available.

In many cases limiting action can be made even more effective by allowing it to take place below the 100 per cent modulation level. This will, of course, introduce distortion into the desired signal during modulation peaks; but, remembering that, in practice, an average modulation level of from 30 to 40 per cent is rarely exceeded, the resulting distortion is quite small as compared to the improvement obtained due to the suppression of interference peaks.

1.8.3.3.2 WAVE TRAPS

A wave trap is a circuit designed to attenuate greatly one frequency, or a very narrow band of frequencies, while passing without appreciable attenuation all other frequencies. Wave traps are usually inserted into the antenna circuit of a receiver to eliminate one particular frequency, which happens to be particularly disturbing. Since the input circuit of a receiver itself acts like a band-pass filter, frequencies far removed from the acceptance band of the receiver will usually be sufficiently attenuated by the receiver itself unless they are extremely strong. Wave traps are, therefore, most frequently employed when the frequency to be suppressed lies just above or just below the frequencies passed by the receiver. In the case of very strong interfering signals, however, the attenuation produced by the receiver may be insufficient, and a wave trap may have to be used even though the interfering frequency is several octaves above or below the frequencies passed by the receiver.

The simplest type of wave trap is an anti-resonant circuit, i.e., a parallel combination of inductance and capacitance as shown in Figure 1.8.3.3.2-A. This circuit has a very high impedance at the anti-resonant frequency and, therefore, attenuates the currents at that frequency. How rapidly the impedance decreases on either side of the anti-resonant frequency depends on the ratio of L/C and on the "Q" of the circuit. This is shown by writing the expression for the impedance of the anti-resonant circuit:

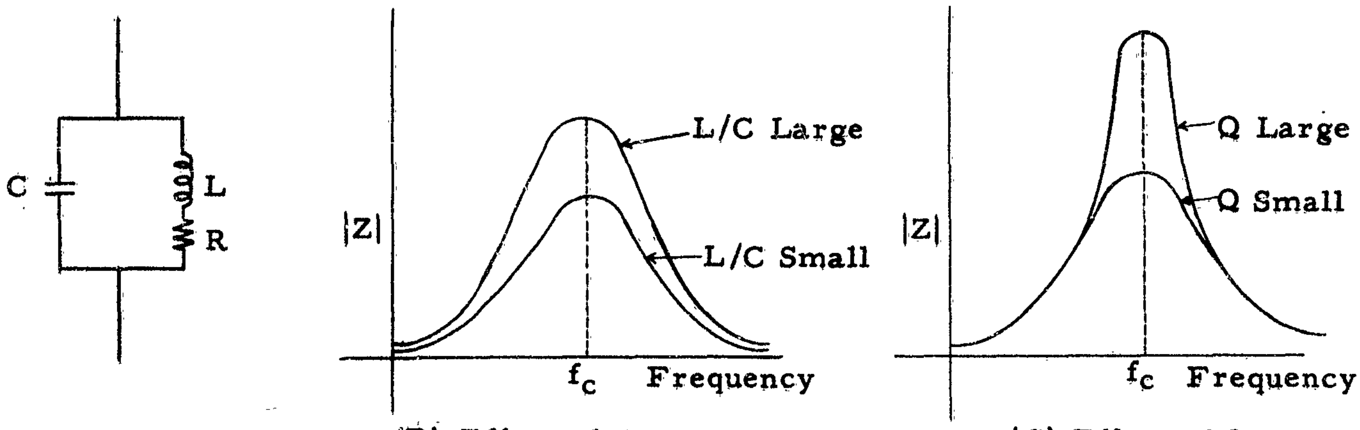
$$|Z| = \left| \frac{R + j\omega L}{j\omega C \left[R + j\left(\omega L - \frac{1}{\omega C}\right)\right]} \right| \tag{40}$$

$$= \frac{f_c^2}{f} 2\pi L \sqrt{\frac{Q^2 + 1}{Q^2 \left[\left(\frac{f_c}{f}\right)^2 - 1\right]^2 + 1}} \tag{41}$$

where f is the frequency and $f_c = 1/(2\pi\sqrt{LC})$ is the approximate anti-resonant frequency. This expression shows that the impedance is decreased at all frequencies if L is decreased with a constant Q and f_c. Since f_c is determined by the product LC, decreasing L with constant f_c means increasing C.

On the other hand, increasing Q with a constant L and f_c has no effect except when f/f_c is close to unity. For it is seen that, if f/f_c is either much larger or much smaller than unity, the impedance is practically independent of Q, provided that Q is larger than about 10 - a condition usually satisfied in practice. But when f/f_c is very close to unity, the impedance becomes practically proportional to Q.

The dependence of the variation of |Z| on Q and on the ratio L/C is illustrated in Figure 1.8.3.3.2. It is seen that, for a wave trap of the type under consideration to have the minimum effect on the transmission of the desired band and to provide the maximum attenuation of the interfering signal, the ratio L/C should be as small as possible and Q should be as large as possible.



(A) Circuit

(B) Effect of the Ratio L/C on the Impedance

(C) Effect of Q on the Impedance

Fig. 1.8.3.3.2 Characteristics of Simple Wave Trap

More detailed information for the design of wave traps will be given in Paragraph 3.1.4.2.

1.8.3.3.3 BLANKING CIRCUITS

In a blanking circuit, the entire receiver is rendered inoperative for the duration of an interfering pulse. Such action can take place before the signal enters the first stage of the receiver, and therefore is not subject to the restrictions imposed by the pulse-lengthening effect of the receiver circuits. Blanking action may be triggered by the interfering pulse itself, in which case a delay line or circuit must be provided for the desired signal so that the receiver is blocked before the undesired pulse can enter it. It may also be triggered by an independent signal in those cases where the arrival of an interfering pulse is known in advance, as for example when the interfering pulse comes from a radar transmitter in the same aircraft. In this case, the triggering pulse may be provided by a signal taken directly from that

radar transmitter. Blanking action is usually provided by a simple amplifier stage that can be biased to cut-off by the suitably amplified trigger pulse.

Blanking circuits are often used when no other methods of protecting the receiver are available, but they lack the simplicity of limiters and wave traps. They are complete units in themselves including amplifiers, trigger pulse amplifiers, and delay circuits. In addition, their action itself may be a source of interference inasmuch as cutting off the carrier periodically is a form of modulation, which will appear in the audio output as noise. This can be overcome by supplying the carrier locally when the blanking circuit punches "holes" into the carrier arriving from the outside, but this increases the complexity of this method still further.

1.8.3.3.4 PHASE CANCELLING CIRCUITS

A phase cancelling circuit is a circuit that allows the transmission of the interfering signal by two different paths. When the two portions of the interfering signal recombine, they are exactly 180° out of phase and cancel. Because the phase shift of any network is usually a function of frequency, phase cancelling networks are not designed for signals containing more than one or at most a few frequencies. They can only be used when the path of entry as well as the nature of the interfering signal are fully known. Therefore, their use is practically restricted to interfering signals from radar transmitters or modulators in the same aircraft, or from similar sources of definite frequency.

The block diagram of a typical phase cancelling circuit is given in Figure 1.8.3.3.4.

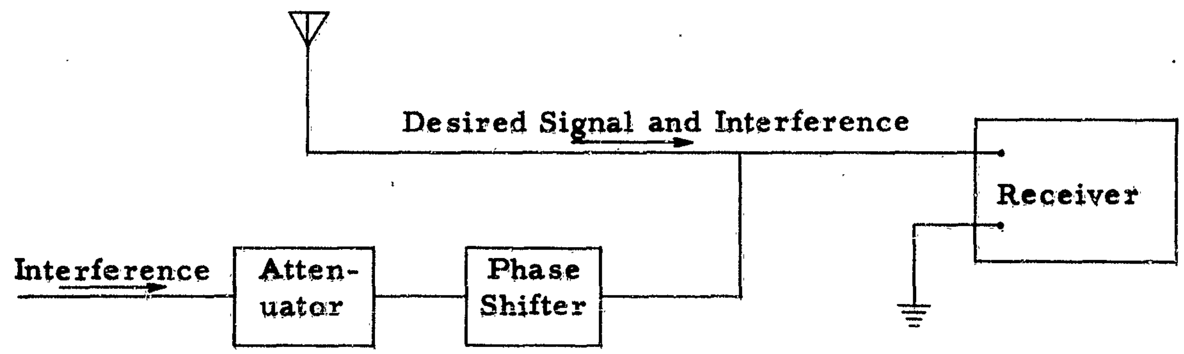


Fig. 1.8.3.3.4 Block Diagram of Phase Cancelling Network

1.8.3.3.5 AUDIO FILTERS

In those cases where the interfering signal contains only a small number of fixed audio frequencies, it is possible to design a filter that eliminates these frequencies and to place it in one of the audio stages of the receiver. For example, when a local radar transmitter causes severe interference at the frequency of its pulse repetition rate, the audio frequencies present in the output of the receiver will be only that repetition frequency and its harmonics. An audio filter can be designed that attenuates the very narrow bands of frequencies about each of these frequencies.

Since the ear is normally quite discriminating against audio signals of different pitch, the desired signals will usually remain intelligible in the presence of even fairly loud interference of definite pitch. An audio filter, in eliminating this interference, will decrease the annoyance caused by the interference, but will not usually appreciably add to the intelligibility of the desired signal. Therefore, and also because of their inherent complexity, these filters are rarely used in existing receivers.

2. THE MEASUREMENT OF INTERFERING SIGNALS

It is obvious that in order to design equipment that is neither a source of, nor susceptible to, radio interference, there must be available adequate means of measuring both the amount of interference generated by, and the degree of susceptibility of, a component, equipment, or complete installation. "Adequate means" here means not only the necessary test equipment, but also standardized test procedures and a set of meaningful limits in terms of easily measured quantities.

As was explained in Paragraph 1.2, the nature of interfering signals, in general, is such that no single quantity can be used for its adequate description. A complete description could be given only in the form of one of two equivalent curves: Either a plot of amplitude as a function of time, or a plot of amplitude (or energy) as a function of frequency. The first is called the wave form, the second the frequency distribution. Neither of these is suitable for quick and easy determination, each requiring a very large number (theoretically an infinite number) of individual measurements.

If the interfering signal consists of a single transient, there is, indeed, no simple method available of obtaining significant information about it by means of a single or a very small number of measurements. But most interfering signals encountered in practice are either periodic, or such that the small differences between samples of the signal taken at different times are not significant. In either case, information about the signal at all times is not required, and attention may be focused on just a short space of time, either a cycle or an arbitrary interval. Furthermore, properties may be defined which can be measured by means of a single measurement, and these properties may be used to characterize the signal. The properties most commonly used are the true average, (rarely used because it is usually zero); the half-cycle average, usually simply called "average"; the root-mean-square; and the peak values of the interfering current or voltage taken either over a period or over an arbitrary time interval.

All interfering signals consist, or may be considered to consist, of a series of pulses. In practice, most such signals may be classified as one of two types: The impulsive type, in which the individual pulses are very short as compared to their average repetition rate, so that a comparatively long period of silence exists between successive pulses; and the random type, often called "white noise", in which the pulses follow one another so closely that the character of the individual pulses is completely lost. There is no sharp boundary between the two, and all intermediate stages are possible. Yet it seems that most interfering signals encountered in practice can definitely be placed in one group or the other.

Interference of the impulsive type will show very high peak values and very low average or root-mean-square values. In fact, on some type of meters reading average or root-mean-square values, an impulsive type of interference may indicate nothing despite having a high nuisance value. Random interference, on the other hand, is best measured by its average or root-mean-square value, while its peak value, which may be only slightly larger than the average and may be subject to considerable random variation, is of no great significance. Thus it is seen that a

meter, or its mode of operation, must be chosen intelligently according to the type of interference to be measured.

Just as the type of meter to be used in measuring generated interference must be chosen according to the type of interference to be measured, so the generator used in susceptibility tests must be chosen according to the type of response of the receiver. Since any type of interference may be present in an aircraft, the susceptibility of a receiver should be tested with a signal of the type which causes the greatest response.

In as much as all "receivers" (as defined in Section 1) must terminate in some kind of indicator or other device capable of responding to the interference, the simplest and most direct check on the presence of interference originating within the aircraft is by routine use of the equipment already installed as a functional part of the airplane. If all electrical components are of proper design and have been correctly installed, interference of any sort will be below a negligible minimum as indicated by headsets, radar-scopes and similar devices. The "General Acceptance Tests" set forth in military specifications discussed in Paragraph 2.1 constitute more than 90% of all checks on aircraft delivered by the vendor to Government agencies. Equipments which do not meet these requirements must go back for rework and then be subjected to much more rigid and detailed tests of a quantitative nature. This is also true in all cases where aircraft, or their components, are being produced under new or original contracts and specifications.

The general acceptance tests have the additional advantage of testing both the efficacy of the sources and the susceptibility of the receivers during the same measuring process. But obviously, such procedure is of little help to the manufacturer who wants information about a single component before it is installed in the aircraft. Hence test instruments and procedures have been set up to enable the manufacturer to evaluate his product in a way which, as much as possible, insures satisfactory performance in the aircraft.

2.1 REVIEW OF MILITARY SPECIFICATIONS

In order to obtain reproducible results, to allow exchange of ideas and information among the various workers in the field, and to insure results that are significant in terms of the ultimate effect of the interference, standard methods of measuring interference must be established. This necessitates the use of standard test equipment and the establishment of standard test procedures. The standards adopted by the Military Procurement Services are laid down in the various military specifications covering radio-interference limits and test procedures. The ultimate purpose of these specifications, which are frequently revised to incorporate the latest developments in the field, is to insure radio-interference-free performance of all electric and electronic installations in the aircraft. The design engineer or other interested personnel should obtain copies of the military specifications of the latest date of issue from the nearest Procurement Service.

In adhering to the specifications, the design engineer must keep in mind the problems of those who are responsible for achieving the ultimate aim: interference-free design. Logistics require that the supply of adequate equipment is not only sufficient, but also accomplished with the minimum complexity of procurement, storing, and distribution. This requires that all aircraft equipment be designed for

the worst conditions likely to be encountered, a policy which results in the greatest flexibility, interchangeability, and ease of relocation of aircraft equipment. For example, interference-generating equipment used only during engine starting (provided it causes no malfunctioning) is not objectionable in land-based aircraft. But the same equipment could not be tolerated in carrier-based aircraft because the interference generated might seriously hamper the communications of the carrier itself. Yet, to keep the total number of parts required by the Military Services at a minimum, the same components should be used by both services if feasible.

The design engineer must also keep in mind that the limits set by the military specifications are reasonable, realistic, and necessary. In all cases, actual limits were imposed only after careful study and the accumulation of a vast amount of experience. Again and again it is found that, as soon as deviations are granted, the final product is unsatisfactory, and much time and money is wasted in order to make the necessary "fixes". Frequently it is found that more time is lost by this procedure than was gained by granting the deviations; and the final product is never as satisfactory as a design that met the specifications from the beginning.

To insure uniformity of procedure and compliance with all details of required tests, whether on component parts or on a completed aircraft assembly, it is required that a government inspector work in cooperation with the contractor. The inspector has general supervision of the test procedure, according to the current issue of the military specifications discussed below, and of the form and extent of the data included in the report. In cases where a contractor desires a deviation from contract specifications or from accepted engineering practice, he must obtain written consent of the procuring service. This service may require detailed drawings and routine tests on the modified device before acceptance.

A brief discussion of the military specifications on radio interference of most interest to the designer of airborne equipment is given below. Since these specifications are frequently revised, the discussion is general and the numerical values given hold only as of the date of issue of this book.

The following specifications are discussed below:

(a) MIL-I-6051. This covers interference compliance tests and general acceptance tests of complete aircraft installations.

(b) MIL-I-6181. This sets limits and prescribes test procedures for both generated interference and susceptibility of aircraft electrical and electronic equipment.

(c) JAN-I-225. This describes procedures for the measurement of generated interference in the range from 0.15 to 20 megacycles for components and complete assemblies.

These three specifications apply to aircraft as of the date of issue of this book. It should be noted, however, that JAN-I-225 is in the process of being revised, and a replacement is expected soon. Also, special specifications for limits and test procedures on aircraft propeller systems are in the process of being written. Finally,

for purchases by the Bureau of Ships, Specification MIL-I-16910, which is also discussed briefly below, should be used in lieu of JAN-I-225.

2.1.1 MILITARY SPECIFICATION MIL-I-6051

This specification is dated 28 March 1950 and supersedes Specification AN-I-24a of 31 July 1947. It covers interference limits and methods of measurement for aircraft radio and electronic installations. All tests under this specification are made on complete aircraft. No test instruments are required other than the normal receiver complement, except that, for quantitative measurements, an output meter is used in addition to the normal output device.

Two types of tests are distinguished: The Interference Compliance Test and the General Acceptance Test. The first requires quantitative measurements and is required of all experimental aircraft, and of the first production models until two consecutive models have passed inspection without rework. The second involves no quantitative measurements and is required of all other models as long as the requirements of the specification are fully met. Any model that fails to pass the tests must be subjected to an Interference Compliance Test, and so must each following aircraft until again two consecutive models have passed without rework.

The requirements are simply that there be no radio interference. Radio interference is defined, as in Section 1, as any disturbance or disturbances which cause an undesirable response or malfunctioning of any electronic equipment. Malfunctioning, in turn, is defined as that type of output which departs from normal, due to interference, in such a manner that the operator or actuating mechanism is unable to differentiate operationally between desired and undesired signals. Finally, undesirable response is defined as a recognizable interruption to normal output which introduces no malfunctioning.

This last definition is clarified by the definition given in the Introduction of this book: Undesirable response is any audible, visible, or otherwise measurable response of a receiver (as defined in Section 1) which is not produced by a desired signal. It is clearly the intent of the specifications to require that no equipment or installation, operating alone or in conjunction with any or all other equipments and installations, shall produce any recognizable response of any receiver.

It should be noted that this specification imposes no requirements on the individual components of aircraft equipments and installations. In this respect the specifications are incomplete since relocation of just one item in the aircraft might mean the difference between meeting and not meeting the requirements. In other words, the fact that a completed aircraft meets the requirements of this specification is no guarantee that each component meets the requirements of Specification MIL-I-6181 (see below). Conversely, strict compliance of all components and individual equipments with the requirements of MIL-I-6181 does not always insure the absence of interference in the completed aircraft since equipments operating in conjunction with one another may produce interference not detectable when either is operating alone. Thus these two specifications complement each other.

2.1.2 MILITARY SPECIFICATION MIL-I-6181

This specification is dated 14 June 1950 and supersedes Specification AN-I-42 of 25 May 1948. It covers interference limits and tests on aircraft electrical and electronic equipment. The tests to be performed are essentially laboratory tests on individual equipments or components.

The basic purpose of this specification is to insure an interference-free design of all electrical and electronic equipment in aircraft. It might be said that the entire purpose of this handbook is simply to aid the designer and manufacturer in meeting the purpose and intent of this specification. Equal emphasis is placed on these two aspects:

(a) Equipment shall be so designed as to generate the least practicable radio interference before interference-suppression components, such as filters or shields, are applied.

(b) Components shall be placed, and circuitry arranged, in such a way as to result in a minimum of undesired coupling.

This specification deals with two kinds of requirements: Those for equipments or components capable of generating interference, and those for receivers capable of being affected by interference.

2.1.2.1 SUSCEPTIBILITY LIMITS AND TESTS

The purpose of the susceptibility tests is to determine the degree to which undesirable signals may gain entrance to, and cause undesirable response or malfunctioning in, the receiver. The requirements are simple: A sine-wave signal of 1000 microvolts (2000 microvolts for frequencies below 150 kilocycles for receivers operating in that range) applied between any external circuit or lead, other than the antenna, and ground shall produce no interference. The leads to be tested include all power and control leads as well as all leads to headphones, scopes, or other indicating devices. When testing the leads connecting the output device, care must be taken that the output device itself is not susceptible to the applied signal. The purpose of this requirements is to insure that the receiver is designed in such a way that no radio-frequency interference can gain entry to the sensitive circuits of the receiver through the output leads, be amplified and detected within the receiver, and then appear as interference in the output. No information about the receiver could be obtained if the output device were affected directly by the test signal.

2.1.2.2 INTERFERENCE LIMITS AND TESTS

The purpose of the interference tests is to determine the amount of interference generated by the equipment or component under test. The requirements are stated separately for conducted and radiated interference. Conducted interference is to be measured only at frequencies from 150 kc to 20 mc, and the maximum allowable voltage, to be measured in accordance with Specification JAN-I-225 (see below), is 50 microvolts, except below 300 kc. Between 150 and 300 kc the limit decreases linearly from 200 to 50 microvolts.

"Radiated interference" is defined in this specification as the interference that is propagated in the form of an electro-magnetic field, including both the radiation and the induction components of the field. Thus it is seen that the term is not used in the strict technical sense, but rather somewhat loosely. This discrepancy was discussed in Paragraph 1.6.4. It is best, in connection with this specification, to think of radiated interference as any interference that is picked up by an antenna or probe not in contact with a lead carrying interfering currents.

Radiated interference must be determined in the frequency range from 150 kc to 1000 mc. Since indications of measured interference depend on the meter used as well as on the pickup device used with the meter, the limits are specified in terms of readings on specific interference meters. The meters specifically recommended are the Ferris Model 32A, Measurements Corporation Model 58, TS-587/U, and AN/URM-28. All of these instruments are discussed below in Paragraph 2.2. If a contractor desires to use any other test instruments, he should obtain written permission from the Procuring Agency. In general it will be required that the meter used be calibrated against one or more of the recommended meters.

It should be noted that this specification requires not only the use of specific meters (or their equivalent), but also specifies the pickup device to be used as well as the position of the functional switch, which, in various meters, allows readings designated variously as "Peak", "Quasi-Peak", or "Field Intensity". The meaning of these designations is discussed in more detail in Paragraph 2.2.1.

2.1.3 MILITARY SPECIFICATION JAN-I-225

This specification, dated 14 June 1945, deals with the test conditions and methods used for making the measurements required by Specification MIL-I-6181. It was written essentially with the Ferris noise meter type 32B in mind, and therefore covers only the range from 150 kc to 20 mc. Its main purpose is to establish a set of standard test conditions which will allow the measurements to be repeated at different times and in different locations with essentially the same results. It therefore specifies the terminating equipment and the arrangement of leads relative to a ground plane of specified dimensions and constructions since it was found that the physical length and arrangement of these leads may affect the readings. Future specifications which will replace JAN-I-225 are expected to deal in more detail with the measurements required at higher frequencies and the many new interference meters which have become available in recent years.

2.1.4 MILITARY SPECIFICATION MIL-I-16910 (SHIPS)

This specification, dated 14 January 1952, is used for Bureau of Ships purchases in lieu of JAN-I-225. It is a very much extended edition of JAN-I-225, covering frequencies from 14 kc to 1000 mc. It deals with all interference meters discussed in Paragraph 2.2, plus the experimental AN/TRM-4 not discussed in this book and two obsolete Navy models.

One of the major changes in this specification as compared with earlier ones is the recognition of the importance of band-width. Limits for interference are given in microvolts per kilocycle band-width for conducted interference, and in microvolts

per meter per kilocycle band-width for radiated interference. An exception is made if the interference is of the continuous-wave type, in which case the older units of microvolts and microvolts per meter are used. A further discussion of the reasons for this change will be found in Paragraph 2.2.

The seventeen meters mentioned in this specification are divided into four groups as follows:

(a) Approved as conforming with the detailed requirements set forth in these specifications.

(b) Approved for a special purpose only.

(c) Not evaluated at the time of classification.

(d) Meters in this group, being existing meters in general use, may be used until conditions permit their replacement with approved types.

The classification of each meter is given with the other pertinent data in Paragraph 2.2.

Since this specification covers a much wider frequency range and a much larger number of test instruments, it provides a much more extensive treatment of the subject than JAN-I-225. For further details, the specification itself must be consulted.

2.2 RADIO INTERFERENCE MEASURING SETS

New instruments are being developed continuously and made available to the manufacturer. Those currently available are discussed below, and the manufacturer should check the Air Force for information regarding newer types whenever the instruments described below do not suffice to make the required measurements on certain equipments.

2.2.1 GENERAL DESIGN CONSIDERATIONS

From the discussion in Section 1.2 and from what has been said in Section 2, it is obvious that the nature of radio interference prohibits its adequate description by any single measurement. Presently available measuring instruments are usually calibrated to read the peak, quasi-peak, effective, or average values of the interference. However, since the electrical disturbances causing interference are complex waves varying greatly in amplitude, phase, and frequency distribution, a single type of measurement can be no more than an arbitrary established standard. Several different measurements of the same interference would be required to adequately describe it.

A radio-frequency type of interference meter measures the RF amplitude of radio interference in much the same manner as a "conventional" field strength meter. It is complete within itself, consisting of a specially designed and constructed radio receiver, and an indicating meter which is connected into the receiver by means of a special circuit. Attention is drawn to the fact that the conventional type

field strength meters are useful only for measuring sinusoidal waves, that is, they do not have the dynamic characteristics necessary for measuring the many types of wave shapes which, as stated above, constitute radio interference. For this reason, any attempt to "fix" or modify a radio receiver in such a way as to make it satisfactory as a general-purpose radio interference meter will be unlikely to succeed.

The audio-frequency type meter is used for the measurement of the effect of radio interference by connecting it across the output of the radio receiver. It is essentially an output meter, provided with special circuits to give quasi-peak, peak, and average readings on a meter which may be calibrated in volts, milliwatts, or decibels. Meters of this type may be used to measure radio-frequency-interference currents and voltages if they are used in conjunction with a calibrated receiver. It is helpful to record, along with the values read on the meter, a qualitative description of the interfering signal obtained by monitoring by ear or eye with headphones and/or an oscilloscope.

The ideal meter for making interference tests should be capable of admitting signals at a very low energy level, amplify them with high fidelity and by known amounts, and present them to a calibrated output meter for measurement. Because of the selectivity and stability of the superheterodyne receiver circuit, it has been used largely as the foundation for all radio-frequency meters of recent design. Attempts to modify regular service receivers to make them function as interference meters have met with only partial success because of inherent limitations. The functions of the essential parts of typical interference meters are shown in the block diagram of Figure 2.2.1.

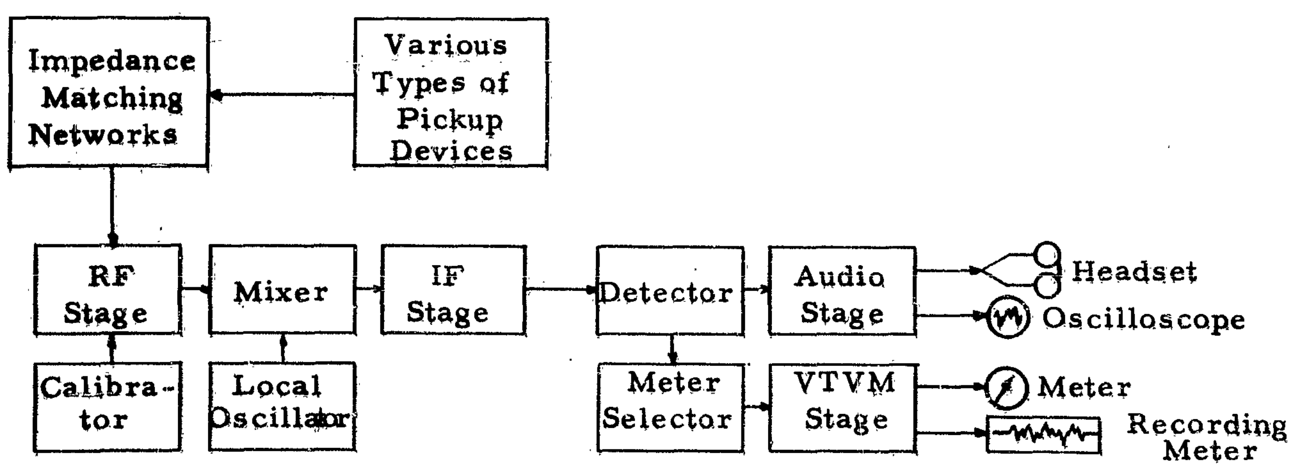


Fig. 2.2.1 Block Diagram of a Typical Radio Interference Meter

In the following paragraphs, each stage is briefly discussed with respect to its function in a radio-interference measuring set. Special emphasis is given to the features which distinguish a radio-interference measuring set from an ordinary receiver: the calibrating device (see Paragraph 2.2.1.4), the detector with its weighting circuits (see Paragraph 2.2.1.7), and other output indicating devices (see Paragraph 2.2.1.9).

2.2.1.1 PICKUP DEVICES

The pickup devices used for radio-interference measurements are either a two-terminal probe, which is used for making measurements of conducted interference, or an antenna, which is used for making measurements of radiated interference in the sense of the interpretation of MIL-I-6181 (see Paragraph 2.1.2.2).

For any kind of pickup device, the first question to be asked is whether it is desired to obtain the maximum indication of the output meter, or whether the most important consideration is to keep the measured circuit undisturbed. In the first case, the pickup device should be designed for an impedance match so that maximum power transfer may take place. In the second case the input impedance of the pickup device should be as high as possible. Since radio-interference measuring sets should give indications of the worst possible conditions, the pickup devices are usually designed for a good impedance match. The measurement of conducted interference is achieved by means of the impedance stabilizing network specified in proposed JAN-I-225 and MIL-I-16910 the main purpose of which is standardization, and minor mismatches are of little consequence. For the measurement of radiated interference, the antenna may be thought of as a device that matches the input impedance of the meter to the impedance of the wave that is being picked up. If the antenna is placed close to the source, which is the condition usually found in radio-interference work, then the difference between the conditions of maximum power transfer and minimum effect on the source still exists. But if the antenna is placed far away from the source (i.e. in the true radiation field as defined in Paragraph 1.6.4), then the difference disappears since the antenna will not react back on the source regardless of its impedance.

The two most common antennas used for the measurement of radiated interference are the rod and the loop antennas. The rod antenna has an impedance of the same order of magnitude as the impedance of free space to plane waves (which is 377 ohms) or higher. It is therefore suitable for plane waves and for high impedance fields found in the vicinity of some sources which produce comparatively large electric and small magnetic fields. Loop antennas have very low impedances and are therefore suitable for low impedance fields such as may be found in the vicinity of some sources which produce comparatively small electric and large magnetic fields. The impedance of a loop antenna may be decreased still further by surrounding the loop with an electrostatic shield, broken in the middle, which decreases the sensitivity for high impedance fields. This feature is sometimes valuable when a high impedance and a low impedance field are present simultaneously and the antenna is to discriminate against the first. It is obvious that in a field whose nature is unknown, measurements must be made both with a rod and with a loop antenna in order to determine the worst possible conditions.

Loop antennas have the property of being strongly directional. The received signal is a maximum if the source lies in the plane of the loop and a minimum if the plane of the loop is perpendicular to the line from the antenna to the source. This property is of great value in locating an unknown source. Special loop antennas with comparatively high impedances are available so that their directional effects may be utilized also for high impedance fields.

With most radio-interference measuring sets, regulation equipment consists

of two vertical rod antennas and several loop antennas of different sizes. The rod antennas have effective electrical heights of one-half and one meter, respectively. They are built in three or four sections which telescope to save space in packing, and have adapter-connectors at the base for attachment to the meter input. The telescoping rods may be conveniently calibrated for lengths corresponding to different input frequencies. Connectors are provided for attaching horizontal wire antennas and for coupling with impedance-matching networks. The loop antennas may be round or rectangular and vary in size from one-half inch to 30 inch diameter or more. The larger loop antennas can be mounted on tripods so that they can be rotated in any direction. Shielded cables are provided to connect these antennas to the meter.

2.2.1.2 IMPEDANCE-MATCHING NETWORKS

Because of the necessity for matching impedances explained above, networks must be provided to match the various types of antennas and probes to the input of the meter. In addition, for the measurement of signals of high energy, an attenuator must be provided in order to avoid overloading of the final circuits. The functions of impedance-matching and attenuation may be conveniently combined in the same network. The impedance-matching network must be suitable for all antennas provided with the meter, and should include provision for a 50 ohm connection for calibration purposes. The attenuation ratios normally provided are 1:1, 10:1, 100:1, 1000:1, and 10,000:1. Instead of providing the attenuation before the first radio-frequency stage by placing the attenuator in the impedance-matching network, a separate attenuator is sometimes placed before the mixer stage.

2.2.1.3 RADIO-FREQUENCY AMPLIFIER STAGE

As in all superheterodyne receivers, the radio-frequency stage consists of a highly selective amplifier with a good gain ratio. When an instrument is designed to cover a wide range of frequencies, for example, 38 mc to 1000 mc in 3 to 5 bands, the tuning condensers must be provided with proper band switches which allow a minimum overlap of 5% between bands. A 2% accuracy on frequency indicated requires high-grade parts and careful calibration.

While commercial practice in the design of broadcast receivers employs one multipurpose tube as RF amplifier and local oscillator, the rigid requirements of measuring instruments requires better performance obtained by using a separate tube in the local oscillator position. One of the advantages is a greater IF rejection ratio and a higher image rejection ratio, both of which are of the order of 60 db.

The major difference between the radio-frequency stage of a normal receiver for amplitude modulated signals and that of a radio-interference meter lies in the effective bandwidth of that stage. For the reception of radio signals amplitude modulated with audio signals, a bandwidth of from 10 to 20 kc is adequate at all frequencies. For a radio-frequency meter, there are two requirements: Firstly, the bandwidth should not exceed a certain fraction of the operating frequency in order to allow the determination of that frequency with fair accuracy. And, secondly, the bandwidth should be large enough to allow interference pulses of short duration to pass through the stage without appreciable lengthening and peak depression. These requirements mean that, at the higher frequencies, the bandwidth must be considerably

larger than 20 kc. Specification MIL-I-16910 requires meters to have an effective bandwidth (defined as the frequency span between half-power points, or between points 6 decibels down from maximum response) of 100 kc for the frequency range from 20 to 1000 mc, and in addition an adjustment for a bandwidth of one megacycle for the frequency range from 200 to 1000 mc. However, it is more important for making accurate measurements that the bandwidth is accurately known than it has a specific predetermined value. Therefore charts giving the exact bandwidth as a function of frequency should be supplied with each meter.

2.2.1.4 CALIBRATOR

Calibration of a radio-interference measuring instrument may be made either with an external signal generator or with a built-in internal source. The source, whether external or internal, may be either a generator of sinusoidal signals or a source of random noise. Any set may be calibrated by means of an external source of either kind provided the proper connectors are available for the proper impedance match and provided the external source furnishes a signal of known magnitude and character. This type of calibration requires no further discussion.

Whether the calibration is to be made with a sinusoidal signal or with a random signal depends on the type of interference to be measured. For accurate results, the signal produced by the calibrating source should resemble the interfering signals as closely as possible. When this is not possible, a certain inaccuracy is necessarily introduced. When a meter has an internal calibrator, considerations of space, weight, and simplicity demand that one or the other type be chosen. About as many sets use random interference sources as use sinusoidal ones. When a sinusoidal oscillator is used, it may either be tunable so that calibration can be made at the test frequency (this is preferable), or it may have a fixed frequency, usually near the middle of the range of the meter. A random-interference source need not be tuned since a single source is usually sufficient to cover the entire range of the set.

Calibration is accomplished by applying the known output voltage of the calibrator to the input terminals of the test set, and then adjusting the gain of the set until the correct output indication is obtained. A convenient source of random voltage is found in the thermal agitation currents in the first tuned circuit of the test set. This voltage has many characteristics that are desirable in a calibration source and it has been used as such in some instances. One important advantage of this source is that it covers the entire radio frequency spectrum almost uniformly and thus requires no tuning. Another advantage is that the magnitude of the voltage generated depends only upon such relatively constant quantities as the impedance, bandwidth, and temperature so that it does not need to be measured. Furthermore, its use involves no extra tubes, parts, or circuit complications. Unfortunately, the magnitude of the voltages obtainable in this way is too low for real usefulness as a calibrator.

A more efficient source of random voltages for calibration purposes is the so-called "shot noise" arising in the plate circuit of a vacuum tube. Shot noise is a result of random variation in the number of electrons emitted from the filament of a thermionic tube. Like thermal agitation, shot noise is uniformly distributed over the entire radio-frequency spectrum, and since its magnitude can be much greater

than that of thermal agitation, it is a much more useful calibrating source. Vacuum diodes are used most frequently for this purpose, but gas diodes, such as neon tubes, may also be used, particularly at the lower frequency ranges.

When the shot noise in a diode is used, the space charge in the tube must be eliminated. Space charge acts as a buffer, smoothing out to some extent the fluctuations which generate the shot noise. Since it is desirable to have the fluctuations as strong as possible, space charge is detrimental in this application. It can be eliminated by lowering the filament temperature and at the same time keeping the plate voltage at a high value, so that all electrons emitted are immediately attracted by the plate and the plate current is limited only by the number of electrons emitted by the cathode. The operation of the tube is then said to be "temperature limited". Under these conditions, there is a simple and definite relationship between the shot noise voltage generated and the direct plate current, which can be read conveniently on a direct-current milliammeter. This type of calibrator is used most frequently for random-interference calibration. It is simple, inexpensive, reliable, and convenient to use, and it has proved to be capable of sufficient accuracy for most measurement work.

Calibration by means of an external standard signal generator, while capable of greater accuracy, requires more equipment and is not as convenient so that its usefulness lies in those cases where the greatest possible accuracy is required. The output of the shot-noise diode is not inherently any less definite or stable than that of a signal generator, but there are some differences in the manner of utilization which account for the greater accuracy of the signal generator in practice. These differences are the following:

(a) When the signal generator is used with a dummy antenna of the correct impedance in the circuit, any variations in the antenna step-up (or step-down) ratio are taken into account. When an internal calibrator is used, a dummy antenna is not usually used and the output of the internal source is connected directly across the first radio-frequency stage. The antenna step-up is taken care of in the calibration curves; variations cannot therefore be compensated for in the calibration process, and a variation in the antenna circuit trimming may result in an error of measurement.

(b) The output of the shot-noise diode is of the order of from 5 to 50 microvolts. Noise contributions of the test receiver itself can effect the resulting output considerably, particularly at the lower frequencies where circuit impedances are large. The superiority of the external signal generator lies in the fact that it can be used at higher voltage levels, and thus the inherent receiver noise may be made negligible by comparison.

(c) The output indication produced by a given shot-noise input depends on the bandwidth of the receiver, so that changes of bandwidth affect the calibration. This last point makes the sinusoidal generator more advantageous only when sinusoidal measurements are to be made. For measurements of common interference signals, the advantage is actually with the random-noise calibrator since variations in the bandwidth affect the calibration and the actual measurement in the same way. Thus, these variations are compensated for by this manner of calibration. As was said before,

greatest accuracy is achieved when the calibrating signal is most nearly like the signal to be measured.

2.2.1.5 MIXER AND LOCAL OSCILLATOR

The mixer and local oscillator stages are substantially the same as found in any good superheterodyne receiver. Though some radio-interference measuring sets utilize the same tube both as mixer and as local oscillator, for the stable operation and high-grade performance required, separate tubes for the two functions are preferable. This procedure also decreases oscillator radiation and improves the image and intermediate-frequency rejection ratio.

2.2.1.6 INTERMEDIATE-FREQUENCY AMPLIFIER STAGES

Special considerations for the intermediate-frequency amplifier stages include the large bandwidth and the wide frequency range covered by some radio-interference measuring sets. Since the gain of a single stage is roughly inversely proportional to its bandwidth, the number of intermediate-frequency stages must be larger than in an ordinary receiver. At least three, and in some cases as many as six, are used in currently available instruments. In order to cover a very wide frequency range, more than one intermediate frequency may have to be used. This also has the advantage that no "gap" need be left for frequencies near the intermediate frequency. For example, a meter covering the range from 150 kc to 20 mc and using an intermediate frequency of 455 kc cannot cover the range from about 400 to 500 kc because these frequencies would require excessively low oscillator frequencies. If the same meter employs two intermediate frequencies, say 455 kc for the low band and 1600 kc for the high band, then special provisions can be made to allow utilization of the second intermediate frequency of 1600 kc for the range from 400 to 500 kc, and thus the entire range may be covered.

Commonly used intermediate frequencies are 12.5 kc, 455 kc, 1600 kc, 12 mc, 30 mc, and 60 mc.

2.2.1.7 DETECTOR AND WEIGHTING CIRCUITS

As in any receiver, the detector functions as a rectifier which demodulates the intermediate-frequency signal and converts it into a pulsating, unidirectional current. However, the weighting circuits associated with the detector are the very heart of the radio-interference measuring set. They determine which feature of the interference shall be indicated by the output meter,

The basic circuit of a diode detector is shown in Figure 2.2.1.7-A. The parallel combination of R and C, across which the output is taken, is called the weighting circuit. When the signal developed in the intermediate-frequency transformer makes the plate of the diode positive with respect to its cathode, a current will flow charging the capacitor. To a first approximation, the presence of the resistor R may be neglected during this charging process, and the voltage appearing on the capacitor is mainly determined by the time constant of the series circuit formed by the plate resistance of the diode and the capacitance C. When the polarity of the voltage reverses, no current can flow through the tube, and the capacitor discharges through the resistance R. The rate of discharge now depends on the time constant

of the loop containing C and R. Thus, by proper choice of R and C, the time constants for charge and discharge may be of any desired value.

If both the charging and the discharging time constants are made comparatively long, the detector circuit is unable to follow any fast variations of the input voltage, such as may be caused by interference pulses, or even a sinusoidal audio modulation. It will produce an indication that is essentially proportional to the carrier strength. When this kind of weighting circuit is used, the output indication is usually labeled "Field Intensity".

When the charging time constant is made much smaller than the discharging time constant, say one millisecond charge and 600 milliseconds discharge, the voltage across the capacitor reaches a value near the peak of the applied voltage and remains at or near this value as long as the peak reoccurs frequently enough to prevent the voltage from decreasing. When the signal is modulated 100 per cent at an audio frequency of 500 cycles per second or higher, this circuit will produce a reading almost twice as high as the "Field Intensity" circuit. The proper name for a circuit of this type is "Quasi-Peak", meaning "almost like peak", but several current instruments label this position "Radio Noise", or even "Peak". The reason for the term "Radio Noise" is that in practice this position very often produces a reading that is approximately indicative of the nuisance value of the interference.

It is practically impossible to obtain a true peak reading for all kinds of pulses by adjusting the parameters of the weighting circuits. To obtain a true peak reading, a so called "slide-back" circuit is made use of. This consists of an adjustable bias on the detector diode as shown in Figure 2.2.1.7-B. This circuit makes the plate of the diode negative with respect to the cathode when no signal is applied. Even with a signal applied, no current can flow, and therefore no output voltage can appear, until the signal produces a voltage large enough to overcome the bias. The procedure is to adjust the bias manually until the aural output indication just disappears, using the highest possible gain of the output stages. At this point the bias is just equal to the peak value of the applied signal. The time constants for this case are usually chosen about the same as for the "Field Intensity" position.

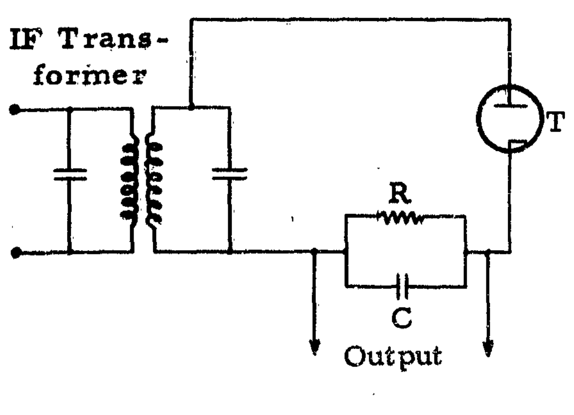


Fig. 2.2.1.7-A Basic Detector Circuit

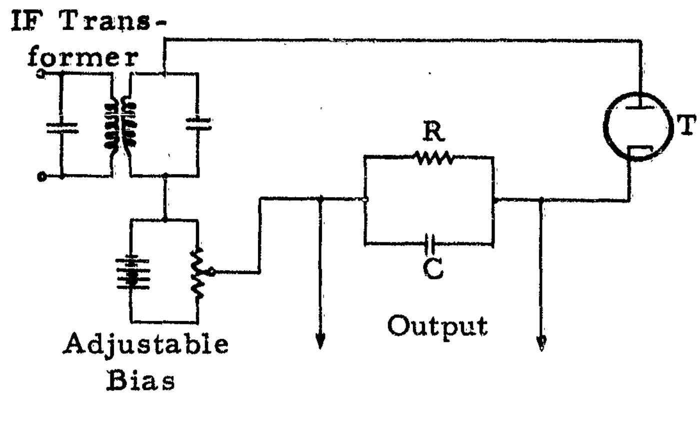


Fig. 2.2.1.7-B Detector with Adjustable Bias

It should be emphasized that the weighting circuits can give information only about the signal as it is presented to them. Since the weighting circuits are placed at the second detector, the carrier strength, quasi-peak, or peak voltage indicated by them refer to the signal after it has passed the radio-frequency and intermediate-frequency amplifier stages. If there is any distortion, pulse lengthening, or peak depression in these stages, the output indication does not give correct information about the input signal to the instrument.

2.2.1.8 AUDIO AMPLIFIER STAGES

The audio signal, passed by the detector and blocking condenser, is boosted by one or two stages of audio-frequency amplification which terminate in a telephone jack through impedances matched with regulation 600 ohm telephone headsets. For purposes of visual study of the signal, terminals for the connection of an oscilloscope are usually provided. These may come directly from the detector or from the output of the first audio amplifier. The amplifier within the oscilloscope is depended upon for obtaining deflections of sufficient amplitude to study wave-form characteristics and to differentiate between audio-frequency modulation and interference.

2.2.1.9 OUTPUT INDICATING DEVICES

To make the maximum use of the potentialities of a radio-interference measuring instrument, the output should be monitored aurally and visually as well as by an indicating meter. Provisions should therefore be made for connecting headphones and cathode-ray oscilloscopes.

The output meter is the final terminal of the amplifiers and metering circuits. It usually consists of a DC milliammeter. Provisions are usually made for a jack into which an outside meter can be plugged, and for another jack to accomodate a recording meter. Since these additional meters are connected in series with the one which is an integral part of the instrument, the jacks contain a matching impedance that is normally in series but is shorted out when replaced by that of the external meter. The service meter may also be provided with multipliers for checking portable batteries used for power-supply.

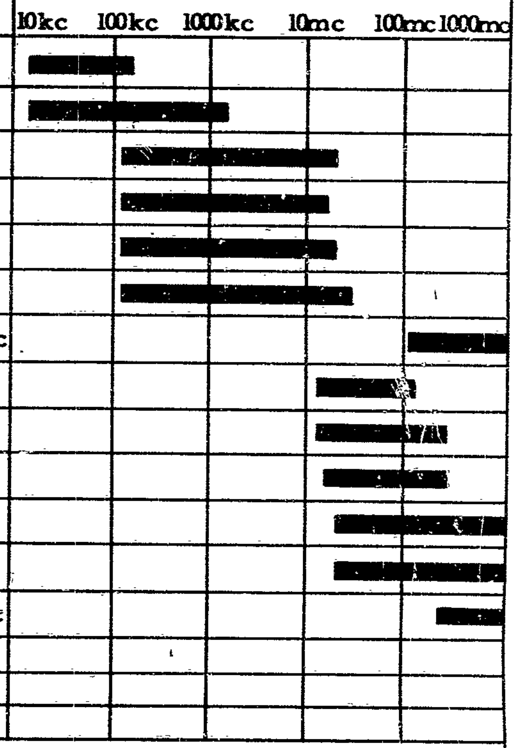
For field intensity work where it becomes important to have a permanent record of the interference pattern around a specified area, or a record of the results obtained from extensive rework of troublesome interference sources, a recording meter may be used. The Army-Navy standard type Milliammeter-Recorder has a range of 0-1 ma and a resistance of 1400 ohms so that it will readily fit into the circuits already described. Its use is limited, however, by the fact that the recording paper is pulled through by a 60 cycles, 115 volts, synchronous motor and can be used only where such a power supply is available. It has changeable gears to provide for different speeds of recording.

2.2.2 DESCRIPTION OF INTERFERENCE METERS

Continued effort has been made in recent years to develop, for military services, interference meters which will give dependable service over the whole list of specifications which may be demanded of them. Instruments which comply in every respect to the specifications which have been drawn to govern interference

FREQUENCY RANGE AND STATUS OF INTERFERENCE MEASURING EQUIPMENTS

EQUIPMENTS	COGNIZANCE	AIR FORCE STATUS	DESIGNATION MIL-I-16910	FREQUENCY RANGE
AN/URM-6	N-(Stoddart)	Approved-ASA-C-63 Spec.	1	14kc to 250kc
FERRIS 64 AB	Comm.	Not Evaluated	3	15kc to 2000kc
FERRIS 64 BC	Comm.	Not Evaluated	3	150kc to 25mc
FERRIS 32 A	Comm.	Acceptable	4	150kc to 20mc
AN/PRM-1	N-(Stoddart)	Approved-ASA-C-63 Spec.	1	150kc to 25mc
AN/URM-3	SC	Acceptable	2	150kc to 40mc
AN/TRM-4	AF(Dev.)	Experimental	3	150mc to 1000mc
MEASUREMENTS 58	AF(Comm.)	Acceptable	4	15mc to 150mc
TS/587-U	N	Acceptable	4	15mc to 400mc
AN/URM-7	SC(Dev.)	Under Evaluation	2	20mc to 400mc
AN/URM-28(AN/APR-4)	AF	Acceptable	4	38mc to 1000mc
AN/URM-29(AN/APR-4)	SC	Acceptable	4	38mc to 1000mc
AN/URM-17	N-(Stoddart)	Under Evaluation	3	375mc to 1000mc



COGNIZANCE KEY
- N — Navy
- AF — Air Force
- SC — Army Signal Corps
- Comm. — Commercial
- AN — Army, Navy, Air Force
- Dev. — Development

See Appendix XIII Descriptive Data Sheets on Interference Test Sets

Fig. 2.2.2

tests are classified as "Approved". Those meters which are generally available and meet specifications except for some minor, though important, details are classified as "Accepted". The Navy specification MIL-I-16910 goes into considerable detail regarding the characteristic of acceptable instruments, and classifies into four principal classes. To assist the reader, the class to which each of the Test Sets described below is noted. Briefly, Class 1 includes those meters which have been designed to meet the standards established and approved by the Armed Services. Class 2 instruments are special purpose and as such do not conform to overall specifications. Class 3 meters have not yet been finally approved by the Armed Services and Class 4 are those in general use, but do not meet most approved standards. Instruments which are in the development stages but not yet ready for production are called "Experimental". The Frequency Range and Status of Interference "Measuring Equipments" as of May 1951 are shown in Figure 2.2.2. More complete data sheets with photographs of Interference Test Sets are given in Appendix XIII.

2.2.2.1 RADIO TEST SET AN/URM-6

This meter has been approved for general use by the U. S. Air Force. It was developed under Navy cognizance and covers a frequency range from 14 to 250 kilocycles. It has attenuator settings corresponding to 0, 20, 40, 50, 60, and 80 db. The calibrator circuit consists of a neon bulb and an amplifier tube together with the necessary tuned circuits. The local oscillator is separate from the mixer, and a very stable beat-frequency oscillator is connected into the grid of the third IF tube. The BFO is used for identification of signals only and must be off while tuning for maximum meter readings. Scale readings are approximately logarithmic. The power input goes through an isolation transformer with filters to keep extraneous power-line noises out of the instrument. It is designed to operate from an AC power source of 105-125 volts, or 210-250 volts, and on frequencies from 50 to 1600 cycles per second. The set is designated as Class 1 in MIL-I-16910.

2.2.2.2 FERRIS MODEL 64 AB

The Ferris Model 64 AB Noise Meter and Field Strength Meter is a contemplated commercial instrument with a frequency range of 15 kc to 2000 kc. It is a later model of Ferris Models 32A and 32B described in Section 2.2.2.4. Details of construction are not available. This set is designated as Class 3 in MIL-I-16910.

2.2.2.3 FERRIS MODEL 64 BC

Ferris Model 64 BC is essentially identical with Model 64 AB except that the frequency range is from 150 kc to 25 mc. This set is designated as Class 3 in MIL-I-16910.

2.2.2.4 FERRIS MODELS 32A AND 32B

Ferris Models 32A and B Radio Noise and Field Strength Meters are essentially similar commercial instruments and are classified as "Accepted" by the Air Force. The range from 150 kc to 20 mc is covered in five bands, 150 kc-350 kc; 550 kc-1550 kc; 1550 kc -4.0 mc; 4.0 mc-10 mc, and 10 mc-20 mc. A gap is left between 350 kc and 550 kc for the intermediate carrier frequency of 455 kc. The circuits

are designed for compactness and simplicity of operation. A single RF stage feeds into a combined mixer and oscillator tube. Two IF stages precede the detector tube. A diode noise oscillator is provided for calibration and a more accurate auxiliary External Calibrator is also available. A logarithmic scale on the output meter is obtained by feeding the RF component of the detector tube back through the control grids of the two IF amplifier stages as is done in conventional AGC circuits. While this noise meter was designed for portability with dry cell batteries, it is also provided with a separate vibrator power pack for 6-volt storage battery service and an AC power pack for 115 volts, 60 cycles supply. This set is designated as Class 4 in MIL-I-16910.

2.2.2.5 RADIO TEST SET AN/PRM-1

Made to Navy specifications in 1949, this meter is on the Air Force approved list. It covers the range from 150 kilocycles to 25 megacycles in seven bands: 150 kc-320 kc; 320 kc-750 kc; 750 kc-1750 kc; 1.75 mc-3.8 mc; 3.8 mc-8 mc; 8 mc-15 mc; 15 mc-25 mc. Intermediate frequencies of 455 kc are used in bands 1, 3, and 4, and 1600 kc in the other four bands. Attenuator settings of X10, X10^2, X10^3, and X10^4 are built into the RF and IF amplifiers. The indicating meter has a two-decade logarithmic scale of 1 to 100 microvolts and an approximately linear scale which runs from 0-40 db. The calibration frequency is fed into the first RF amplifier. The mixer consists of two tuned stages, the mixer and the amplifier, and the local oscillator. Four IF amplifier stages and two stages of AF amplification are used. A balanced bridge circuit with two pentodes forms the VTVM circuit, the unbalance being created by potentials from the receiver and AGC outputs. The power requirements for this meter are 1.1 volts at .86 amperes for the "A" battery and 75 volts, 30 milliamperes for the "B" battery, all of which can be provided by means of regulation type dry cells. The meter is also equipped for obtaining power from 115 or 230 volts AC, 50 to 1600 cycles per second. For 115 volts, 60 cycles power supply, constant voltage is maintained through a regulating transformer with two secondaries. One has high leakage-reactance; and the other, shunted by a condenser, is a resonant loop. The 1.1 volt filament supply is obtained through a step-down transformer and a bridge-connected selenium rectifier with filter. The 75 volt plate potential comes directly from the secondary of the regulating transformer, before rectification. The rectifier is a selenium bridge and close regulation is obtained by means of an OA-3/VR 75 tube. For frequencies other than 60 cycles per second where the resonant loop cannot be used without modification, a thermo-regulator tube serves as a voltage regulator. An autotransformer makes provision for power-supplies of 230 volts. This set is designated as Class 1 in MIL-I-16910.

2.2.2.6 RADIO INTERFERENCE TEST SET AN/URM-3

Radio Interference Measuring Set AN/URM-3, covering the frequency ranges of 0.15 to 0.4 mc, and 1.6 to 40 mc, was designed and developed by the Signal Corps Engineering Laboratories primarily for the measurement of "broadband" interference and represents a new concept in the measurement of broadband interference. It is approved and in use by the Army for interference measurements. The equipment consists basically of a superheterodyne receiver and a calibrated impulse noise generator generating pulses exhibiting a stable and uniform spectrum up to 40 mc, the peak value of which is adjustable to known values.

Interference measurements are made in terms of the output of the calibrated impulse noise generator by aural comparison at threshold level with the "unknown" interference. This method provides an accurate measurement of the peak value of the impulse interference in terms of microvolts per unit bandwidth and is independent of the gain and bandwidth of the measurement receiver. The method of injecting the output of the calibrated impulse generator into the receiver input is such that interference measurement is independent of antenna impedance.

A nine-foot rod antenna is provided for radiation measurements for increased sensitivity. It is not calibrated for effective height because of the manner in which it is used for interference measurements (in close proximity to the equipment being tested and in a non-uniform field over its length). In measurements under these conditions such a calibration including the term "per meter" would be invalid.

Accessories include magnetic and electric field probes for exploration purposes, and matching and coupling networks for use of the test set as a two-terminal RF microvoltmeter. Plug-in power supplies are provided for 115 volt AC and 12 and 24 volt DC operation. This set is designated as Class 2 in MIL-I-16910.

2.2.2.7 MEASUREMENTS MODEL 58

The meter designated Measurements Model 58 is produced by the Measurements Corporation of Boonton, New Jersey. It is one of the earlier models and has been in regular production with some improvements since 1946. It is officially classified as "Accepted". The frequency range of Model 58 is from 15 mc to 150 mc in five bands: 15 mc-24 mc; 24 mc-39 mc; 38 mc-62 mc; 60 mc-100 mc; 98 mc-150 mc. Four decade steps of attenuation, X1, X10, $X10^2$, and $X10^3$, are inserted between the antenna input and the first RF stage. It will respond to a wide variety of signals including AM, FM, pulse modulated, and continuous wave. It can also measure TV carriers at the blanking level. It makes use of one RF stage and three IF stages with an intermediate frequency of 12 mc. A separate tube is employed as a local oscillator and another as a shot-noise diode. The output of the detector tube excites the grid of an amplifier tube which is in one arm of a bridge circuit. The indicating microvolt meter serves as a balance indicator across the bridge arms with readings that are semi-logarithmic. An audio-frequency stage is provided beyond the detector and suitable jacks are provided for connecting an oscillograph, or a recording milliammeter. The power supply follows conventional lines. There is no voltage regulation in the input of the 115 volt transformers for 60 cycle operation but the rectified output is fully regulated. When the primary power source is a 6 volt battery at 12 amperes a vibrator and step-up transformer provide the required 115 volts AC. This set is designated as Class 4 in MIL-I-16910.

2.2.2.8 NOISE-FIELD INTENSITY METER TS-587/U AND TS-587A/U

Noise-Field Intensity Meter TS-587/U or TS-587A/U is also listed as NMA-5. It is acceptable only until a replacement is available. This meter operates from 15 mc to 400 mc in four bands. There are three low frequency bands: 15 mc-31 mc; 29 mc-64 mc; 60 mc-125 mc and one high frequency band: 100 mc-400 mc. The intermediate frequency employed for the three lower ranges is 12 mc and 30 mc is used for the upper band. Its construction is unique in that the IF amplifiers for the two intermediate frequencies are completely independent of each other as to tube

complements and transformers. This improves the tuning, selectivity, and stability of the circuits but adds to the weight and complexity of the instrument by requiring more tubes and parts. The attenuator has values of X1, X10, $X10^2$, $X10^3$. When set at X1 the output meter has a range of 0-100 microvolts. A decibel scale is also provided. Another feature of the TS-587/U meter which adds to stability is the use of a grounded-grid push-pull amplifier in the 100-400 mc RF stage. This instrument employs four stages of IF amplification in each channel and two AF stages beyond the detector. The metering circuit for the VTVM contains two amplifier tubes in a balanced circuit with the indicating meter connected between the plates. The control grid of one tube is connected to ground, unbalance being produced by the output of the detector fed into the grid of the other tube. The 60 cycle, 115 volt power input contains a line filter but has no voltage regulation in the input. The rectifier and filter for +B supply is conventional with electronic voltage regulation added. A separate rectifier, through an RC network, provides 45 volts DC for the plates of the balanced tubes in the metering circuit. This set is designated as Class 4 in MIL-I-16910.

2.2.2.9 RADIO INTERFERENCE MEASURING SET AN/URM-7

Radio Interference Measuring Set AN/URM-7 is a radio interference and field intensity meter developed by the Signal Corps as an engineering instrument primarily for the measurement of broadband interference, although incorporating facilities for CW interference and field intensity measurements. Inasmuch as broadband interference has spectrum properties, the Radio Interference Measuring Set is designed to measure the peak value of such interference in terms of microvolts per megacycle bandwidth over the range 20 to 400 megacycles. To accomplish this type of measurement, the instrument incorporates a noise reference standard, the output of which is calibrated in terms of microvolts per unit bandwidth.

The noise reference standard is an impulse generator which generates impulses of the order of 5×10^{-10} seconds in duration and as a result these impulses exhibit uniform stable spectra out to at least 400 megacycles which is the highest frequency to which the Test Set can be tuned. The impulse generator output is injected into the input circuit of the tuner portion of the Test Set in such a manner as to permit measurement of open-circuited antenna terminal voltage on a per megacycles basis.

The tuner portion of the Test Set utilizes a superheterodyne circuit. The frequency range 20 to 400 megacycles is covered by means of two plug-in type RF heads, the first of which tunes from 20 to 200 mc in two bands and incorporates a 10.7 megacycles IF strip, and the second head tunes continuously from 200 to 400 megacycles and incorporates a 30 megacycle IF strip. The visual output indicator is a peak-reading vacuum tube voltmeter with a logarithmic scale calibrated in microvolts and a linear decibel scale calibrated in terms of decibels above one microvolt per megacycle. The logarithmic scale characteristic is achieved through tapered pole-pieces in the indicating movement, which eliminates the necessity of using automatic gain control with its inherent undesirable effects on noise measurement. The VTVM time constants are such as to permit peak measurement, within 10 percent, at repetition rates down to ten pulses per second. The Test Set can be operated directly from 110 volt AC power and is equipped with a converter unit to permit 24 volt DC operation.

The Test Set is to be equipped with two antennas, a dipole type which can be resonated at each test frequency, and a broadband antenna which requires less frequent adjustment. The dipole antenna is used for field intensity measuring applications where the effective height must be known. The broadband antenna is used in suppression test applications where the antenna must be placed close to the source of interference, e.g., testing an automotive vehicle for interference emanation. The Test Set is equipped with probes for conducting exploratory interference tests and coupling-networks to permit use of the instrument as a two-terminal noise microvoltmeter. This set is designated as Class 2 in MIL-I-16910.

2.2.2.10 RECEIVING EQUIPMENT AN/URM-28

Receiving Equipment AN/URM-28 is also listed as AN/APR-4. It is designed to measure frequency, modulation, and signal strength of radio and radar signals within the range of 38 to 1000 mc.

Since this instrument was constructed in 1944 as a high frequency receiver rather than as a noise meter, it does not have a noise calibrator nor does it have the weighting circuits now commonly used for peak and quasi-peak measurements. It has a complete set of antennas for the various frequencies in its range and is well adapted to field intensity measurements. Within these limitations it is classified "Accepted" by the U. S. Air Force. Three tuning units, quickly interchangeable, cover the frequency ranges as follows:

TN-16/APR-4	38 mc - 94 mc
TN-17/APR-4	74 mc - 320 mc
TN-18/APR-4	300 mc - 1000 mc

A beat-frequency oscillator provides a 1000 cycles tone through the IF stages. Two stages of video amplification terminate in connections for the attachment of radarscope equipment. The power input transformer has windings for 80 volts and for 115 volts and is compensated for frequencies from 60 to 2600 cycles. Instead of the conventional band switches, tuning across the frequency range is accomplished by means of an "Autosweep-Manual Switch" which becomes "Auto Sweep" when a motor-driven mechanism is turned on. The intermediate frequency is 30 mc through five IF stages. In some models a crystal detector is used in place of a triode in the mixer position, while the local oscillator is separate. This set is designated as Class 4 by MIL-I-16910.

2.2.2.11 RADIO-INTERFERENCE MEASURING SET AN/URM-29

Radio-Interference Measuring Set AN/URM-29 is Receiving Equipment AN/APR-4, a radar search receiver, modified in a number of respects to render it suitable for interference measurements over the frequency range of 38 to 1000 megacycles. The principle modification to the receiver accomplish the following objectives: (1) peak response of the receiver at its second detector to interference impulses at its input is determinable by an aural indication which is independent of pulse repetition rate and the hearing acuity of the operator; (2) the stability of the receiver is considerably improved over the unmodified equipment; and (3) the spurious responses in the range of 300 to 1000 mc are greatly reduced.

This equipment measures interference directly in terms of db above one microvolt per megacycle which is readily convertible to microvolts per megacycle or per kilocycle. It does not incorporate an impulse noise calibrator but is calibrated in the laboratory by means of a calibrated standard impulse generator of known spectral intensity in terms of microvolts per unit bandwidth. Measurement of the peak value of impulse type interference is made by an aural slide-back method.

The antenna used with this equipment is a special broadband antenna, designed by the Signal Corps Engineering Laboratories, which requires changes of its configuration only at three points over the frequency range of 38 to 1000 mc. The antenna is used in specified positions, empirically determined, close to the equipment being tested for interference, and is not calibrated for effective height for determination of field strength in terms of microvolts per megacycle per meter because such a measurement is meaningless when taken as close to the equipment as is necessary for such tests. This set is designated as Class 4 by MIL-I-16910.

2.2.2.12 MEASUREMENTS DEVELOPMENT MODEL (AN/TRM-4)

This meter is under development by the USAF and is yet in the experimental stage. It is designed to have a frequency range of 150 mc to 1000 mc. This set is designated as Class 3 by MIL-I-16910.

2.2.2.13 RADIO TEST SET AN/URM-17

Radio Test Set AN/URM-17 is a portable, single band, superheterodyne type radio interference and field intensity meter. It may be used to measure radiated or conducted radio interference in the frequency range of 375 to 1000 megacycles. As a radio frequency voltmeter it will measure voltages from 10 microvolts to 10 volts. As a field intensity meter it will measure from 100 microvolts per meter to 100 volts per meter. The intermediate frequency is 60 megacycles. The effective bandwidth varies from 1.8 megacycles at a signal frequency of 1000 megacycles to 1.0 megacycles at 370 megacycles signal frequency. Charts are provided with the equipment showing effective bandwidth as a function of frequency of the measured signal. The attenuator control, graduated in X10 steps provides for attenuations from X10 to X100,000. It is designed to operate from an AC power source of 105 to 125 volts, or 210 to 250 volts, between 50 and 1600 cycles per second. This set is designated as Class 3 by MIL-I-16910.

2.2.3 SUMMARY OF INTERFERENCE METERS

The currently available interference meters may be divided roughly into three groups according to their frequency ranges. These are:

(a) Low frequency meters (14 to 250 kc). Only the AN/URM-6 is discussed in this group.

(b) Medium and high frequency meters (150 kc to 20 or 40 mc). The Ferris Model 32A, the AN/PRM-1, and the AN/URM-3 are discussed in this group.

(c) Very-high frequency meters (15 to 400 mc). The Measurements Corporation Model 58 and the TS-587/U are discussed in this group. Ultra-high

frequency meters (150 to 1000 mc). Only the AN/URM-17 is discussed in this group.

In addition there are the AN/URM-28 and AN/URM-29, which cover the range from 38 to 1000 mc, but these are not true radio-interference measuring sets, being converted AN/APR-4 receivers.

In the low frequency range, the AN/URM-6 is approved by the Air Force and also designated as Class 1 by MIL-I-16910 (see Paragraph 2.1.4). This meter has satisfactory sensitivity and bandwidth and has the three types of weighting circuits properly designated as "Field Intensity" and "Quasi-Peak", and "Peak".

In the medium and high frequency ranges, the AN/PRM-1 is the latest general-purpose instrument. It is approved by the Air Force and designated as Class 1 by MIL-I-16910. It has satisfactory sensitivity and bandwidth and has the three types of weighting circuits with the proper labels. The Ferris 32A, which is widely used at this time, does not have satisfactory bandwidth and has no provision for a true peak reading. The position labeled "Radio Noise" actually gives a quasi-peak reading. The AN/URM-3 is a special-purpose instrument developed by the Signal Corps for making measurements on impulse-type interference. This meter has the unique feature of having incorporated a standard pulse generator with adjustable pulse-repetition rate for calibration purposes. This makes it possible to calibrate this meter by means of a signal that is practically identical with the signal to be measured. Hence no weighting circuit is required in this meter.

In the very-high frequency range, both the Measurements Model 58 and the TS-587/U are acceptable to the Air Force, and both are Class 4 per MIL-I-16910. The TS-587/U covers a somewhat larger frequency range, going up to 400 mc while the Model 58 only goes up to 150 mc. The Model 58 does not have a true peak indication though one of the control positions is labeled "Peak". This, however, is really a quasi-peak indication. The TS-587/U gives readings in the "Field-Intensity" position, that are averages over about 500 milliseconds. In the "Quasi-Peak" position, the charge time constant is about one millisecond and the discharge time constant is about 300 milliseconds. Thus the quasi-peak reading is not as close to the actual peak for low pulse-repetition rates as it would be for a 600 millisecond discharge. The Model 58 may be equipped to obtain peak readings by slide-back method.

In the ultra-high frequency range, the AN/URM-17 has not yet been evaluated, and no comparisons can be made. It does, however, have satisfactory sensitivity and bandwidth and has the three types of weighting circuits properly labeled.

2.2.4 SELECTION OF TEST LOCATIONS

Most radio-interference tests made on components are performed in a laboratory. In order to guard against ambient-interference, it is necessary to perform the tests in a shielded enclosure. A detailed discussion of the design and construction of shielded rooms will be found in Appendix VIII. In a well constructed shielded room, the most sensitive instrument should not be able to pick up any signal whose origin is outside of the room.

Occasionally it is necessary to make radio-interference measurements at a

designated location other than in a shielded room, and it is then essential to consider the effect of the surrounding objects. If, however, the location can be selected, the observer must know what constitutes a good site for a given set of conditions.

In selecting a site for radio-interference measurements, the following considerations should be kept in mind: The radio frequency voltage induced in the receiving antenna is proportional to the intensity of the electromagnetic field in the space occupied by the antenna. The intensity at this point will depend upon the radiated power; the frequency of the received signal; the distance from the transmitter; the attenuation over the path between the transmitter and the receiver; reflections and reradiations from nearby conductors, such as power lines, wire fences, and steel buildings; absorption by trees; and the effects of hills, gullies, or cliffs.

The ideal site would be on open, flat terrain at a considerable distance (1000 feet or more) from buildings, electric lines, fences, etc. Even buried cables can cause serious effects at low frequencies; since the lower the frequency, the greater the depth of ground penetration. Ideal sites are rare in the more populated sections of the country; therefore, it is good practice to check a proposed location by taking measurements on the desired signal at several points in the vicinity. If the same value of field intensity is obtained at each of the points, any one may be considered satisfactory. If it is necessary to use an unsatisfactory site, a series of readings should be recorded at a number of different points in the neighborhood of the selected site, and copious notes on the site conditions should be appended to the recorded data.

There are no fixed rules to govern the minimum distance between the field intensity measuring equipment and the nearest wire lines or other disturbing objects because possible resonances are unpredictable. Some sites have been found to be satisfactory for low-frequency measurements where the distance was as close as 100 feet, but a check of the location should always be made if any doubts exist. Another consideration in site selections is the possible presence of local electrical interference sources which may make field intensity measurements difficult.

3. APPLICATION OF THEORY TO PRACTICE

In this section the principles developed in Section I will be applied to specific radio interference problems encountered in component and system design. A representative group of components and systems will be discussed in detail from the standpoint of design techniques useful in producing interference-free aircraft. For the most part, the systems to be considered are those which might be encountered in a representative bomber-type aircraft.

Individual components must be designed both for minimum generation of, and for minimum susceptibility to, radio interference. It is important to keep in mind from the very beginning that the component must perform satisfactorily not only alone, but also in conjunction with other components as part of a system. Similarly, each system must be designed for interference-free operation not only when operating alone, but also when operating simultaneously with all other systems and their components in the aircraft. Figure 3 shows the operating periods of a number of electrical and electronic systems in a typical military aircraft. This chart indicates clearly the necessity of considering simultaneous operation of practically all systems when dealing with radio-interference problems.

Filtering, shielding, bonding, etc., which have more or less universal application, will be dealt with separately. Paragraphs are also included which deal with minor components such as switches, brushes, vibrators, etc. Equipments and systems will be broken down for most effective presentation of the suppression techniques involved.

3.1 GENERAL CONSIDERATIONS FOR INTERFERENCE-FREE OPERATION

In the design of equipment and in the layout of systems for interference-free operation the principles set forth in Section I may be used as a guide. Practical considerations, however, will greatly influence the designer's decision as to which suppression technique will be applied to each interference problem as it arises. Factors other than technical which may influence the designer are the following:

(a) Weight. It is of the utmost importance to apply techniques which effectively provide the highest ratio of attenuation per unit of weight since any added weight to the aircraft reduces its payload. It is the designer's responsibility, therefore, to study the interference problem in order to insure minimum weight consistent with interference levels which can be tolerated. The most effective suppression techniques are available during the initial design stages of components or systems. Once a design is completed, the designer is limited to techniques which are less effective, often resulting in corrective modifications and additions involving added weight. In many cases proper design may result in elimination, rather than suppression, of interference. In any event, it is

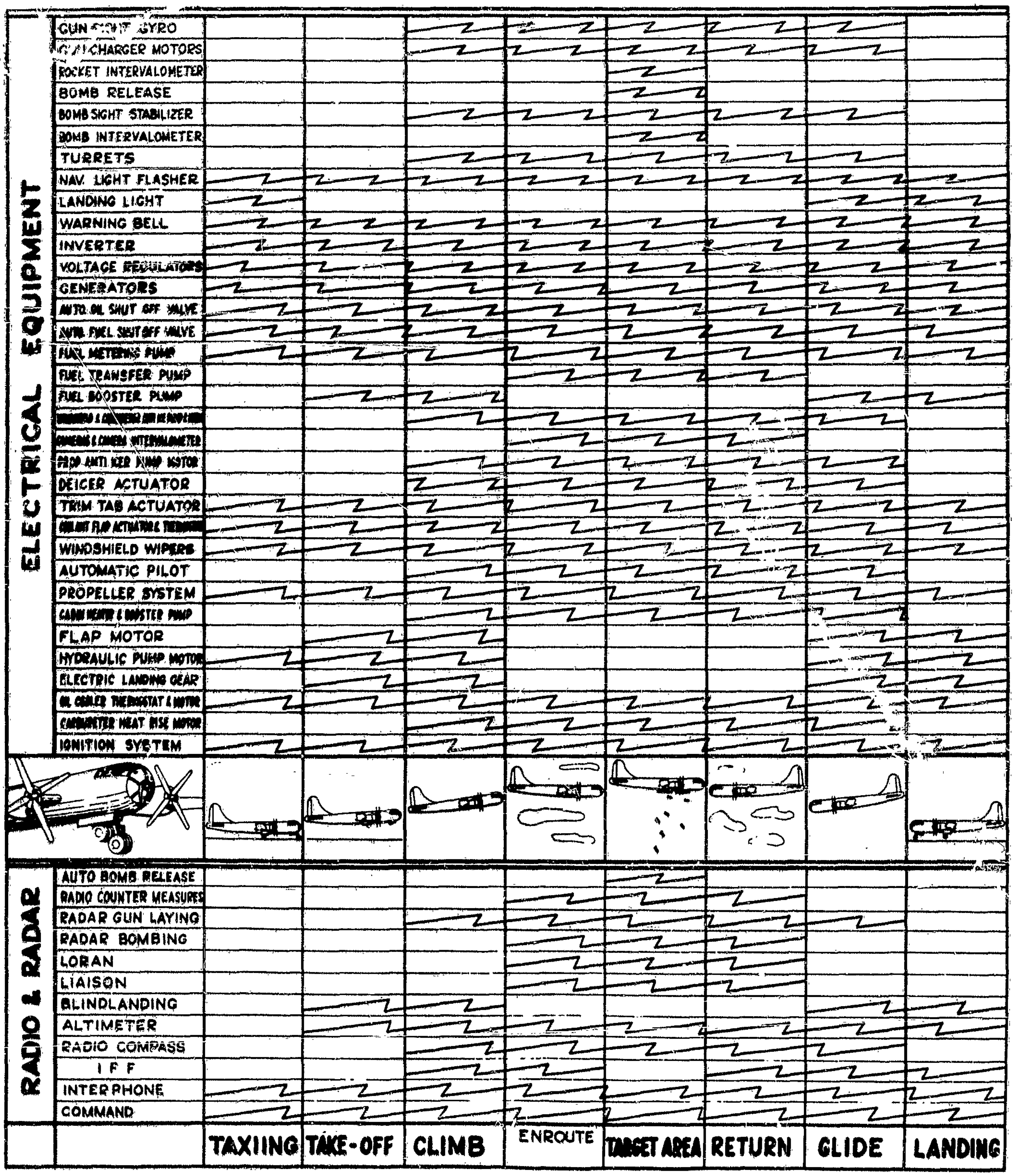


Fig. 3 Operational Periods of Radio, Radar and Electrical Equipment During a Typical Combat Mission

obvious that the best control of radio interference will result if it is considered early in every stage of component or system design.

(b) Space. The use of proper design techniques in designing components and systems from the standpoint of radio interference will minimize the size of components, the number of filter networks required, and the shields required in installations. This is particularly important in modern aircraft where high performance is essential. To achieve high performance all dimensions are critical and components and systems must be designed to fit the space available. This does not mean that radio interference can be ignored. Rather it requires better design practices in all phases of design.

(c) Materials. Often the designer is influenced by availability of materials in the design of suppression devices to attenuate or eliminate radio interference. For example, inductances obtainable for filters in high current circuits are limited by the direct current resistance of the winding. Filter capacitors, especially those of high values, may change considerably at high temperatures. As a result of the use of materials of construction which will not "stay put" under operating conditions, the attenuation of radio interference may be permanently impaired, or if these changes vary rapidly with time, the change of impedance will actually result in additional sources of radio interference. These and similar factors must be borne in mind throughout the design of components and systems if the design is to be interference-free under all operating conditions.

Practical design must take into consideration all of the above, and the best compromise must be used. In addition to these factors the designer must choose the most efficient methods of eliminating or suppressing radio interference.

It may be shown that there are three places in which interference can be controlled: (1) At or in the source; (2) Along the coupling path; (3) At or in the receiver affected. Depending on the nature of the interference, the transmission path between source and receiver, and the receiver characteristics, the designer may find that there is a choice of several alternative solutions to control the interference. This permits the designer to choose the most effective method and to minimize the weight, size, and use of critical materials. The first and third are areas in which the designers of components and equipments have control. The designers of airplanes and aircraft systems control the second.

Practically, a cooperative effort by all designers in all three areas is the only solution to the problem of radio interference. Since it is necessary to consider weight, size, and materials in the functional design of equipment as well as the problems of radio interference, the designer will choose a compromise which will permit satisfactory aircraft operation. This implies radio noise suppression to a degree rather than complete elimination. This again is a practical compromise since there is no such thing as a realizable filter with infinite attenuation or a receiver with zero susceptibility to interference. Specifications (see Paragraph 2.1) have been written covering interference limits and receiver susceptibility. The limits were set as high as is believed to be tolerable in order to achieve what is

considered interference-free aircraft operation.

These limits should never be exceeded and the designer should strive for as nearly interference-free operation as is practically possible. While the present limits seem to be satisfactory, they may at some future time be made more severe as the multiplicity of equipments in aircraft increases. It is conceivable also that a particular installation of equipments, each of which is within specified limits, may interfere with each other because of some installation peculiarity.

In the design of any equipment, system, or installation, the designer should keep in mind the fact that it must operate satisfactorily in conjunction with all the other equipment which may be in the airplane and which may operate at the same time. If this is considered at the inception of the design, it is possible to make most effective use of techniques for interference control. In many cases a study of various approaches to a problem yields one in which no appreciable weight need be added to achieve the required degree of suppression. In fact, proper design may even decrease the weight originally considered necessary, simply by careful routing of wires, correctly orienting parts or assemblies, shielding, or using correct design values in filter networks.

3.1.1 FILTERS AND OTHER SUPPRESSION NETWORKS

The purpose of filters or other suppression networks is to attenuate interfering signals conducted from interference sources or into interference-susceptible equipment. The design of filters will depend on many factors such as the character of the interference, whether the filter is to be inserted at the source or the receiver, and the number of equipments affected. Other things being equal, it is generally preferable to eliminate or suppress interference at the source.

Radio interference filters may require unique characteristics due to the complexity of present day aircraft installations of electronic equipment. For example, a filter may be required with large attenuation from 150 kc to 1000 mc, a range that cannot usually be covered with a single-section filter. Filter design is usually based on a knowledge of the terminating impedances. In many cases, complete information may not be available over the entire frequency range that must be considered in radio-interference problems. In general, a fair approximation of the impedance-versus-frequency characteristic of the circuit at the frequencies to be passed suffices to work out the network parameters.

Three types of suppression networks are most important for the suppression of conducted radio interference. The "brute-force filters", consisting of a single capacitor, a single inductor, or a simple L-section, are used mostly in direct current circuits; power-line filters are used in 60, 400, or 800-cycle alternating current circuits either at the output of a source of interference or at the input of a receiver; and harmonic-suppression filters are used in the output of transmitters. These types are treated in the next three paragraphs. Following them, the design of individual elements for suppression networks are discussed in detail.

3.1.1.1 "BRUTE-FORCE FILTERS"

The simplest type of suppression network is a single capacitor connected from

a conductor carrying interfering currents to ground. Such a capacitor offers a low impedance to ground to the interfering currents while offering infinite impedance to direct currents. It is very effective as long as the capacitor behaves like a capacitance, i.e., as long as the inductance of the capacitor leads are negligible. The problems associated with this network are those of capacitor design, to be discussed in Paragraphs 3.1.1.4 and 3.1.1.5, and those of proper installation to be discussed here.

The capacitor must be installed as close to the actual point of interference generation as is physically possible. For equipments enclosed in a case or shield, this always means that the capacitor must be located within the case or shield. In the case of direct current motors or generators, the capacitor should be connected directly across the brushes. Grounding of the capacitor directly to the negative brush is preferable to grounding it to the frame or case. An exception to this occurs in the case of feed-through capacitors used when the interference source is completely shielded (see Paragraph 3.1.1.5).

It is extremely important to install the capacitor in such a way that the connecting leads are as short as possible. Any lead wire has inductance, and at a certain frequency this inductance resonates with the capacitance. Above the resonant frequency, the effectiveness of the capacitor for the reduction of radio interference decreases very rapidly. This is shown by Figure 3.1.1.1-A, which gives the insertion loss as a function of frequency for a 4 μf condenser, installed in a 28-volt direct current line operating into a 10-ohm resistive load. Curve A is plotted for a condenser connected by means of a lead wire 4 inches long; Curve B is plotted for a condenser with connections made directly at the terminals. In general, the insertion loss in decibels of an ideal capacitor, connected across a load Z_R which is fed by a generator of internal impedance Z_s, is given by the expression:

$$\text{Insertion Loss} = 20 \log \left| 1 + j2\pi f C \frac{Z_s Z_R}{Z_s + Z_R} \right| \tag{3-1}$$

where f is the frequency in megacycles per second, C the capacitance in microfarads, and the impedances are in ohms. If both Z_s and Z_R are pure resistances, i.e., $Z_s = R_s$ and $Z_R = R_R$, then this reduces to:

$$\text{Insertion Loss} = 10 \log \left[1 + \left(2\pi f C \frac{R_s R_R}{R_s + R_R} \right)^2 \right] \tag{3-2}$$

Figure 3.1.1.1-B gives the resonant frequencies of capacitors of various sizes plotted versus the total length of both leads. These curves must be considered approximate since the geometrical arrangements of the leads and the internal construction of the capacitor itself affects the resonant frequency to a considerable extent. The curves shown are based on experimental data, but similar curves (actually straight lines) would be obtained by assuming that the leads have an inductance of 25 milli-microhenries per inch. Such an inductance would exist if the leads are run parallel to a ground plane and three wire diameters away from it.

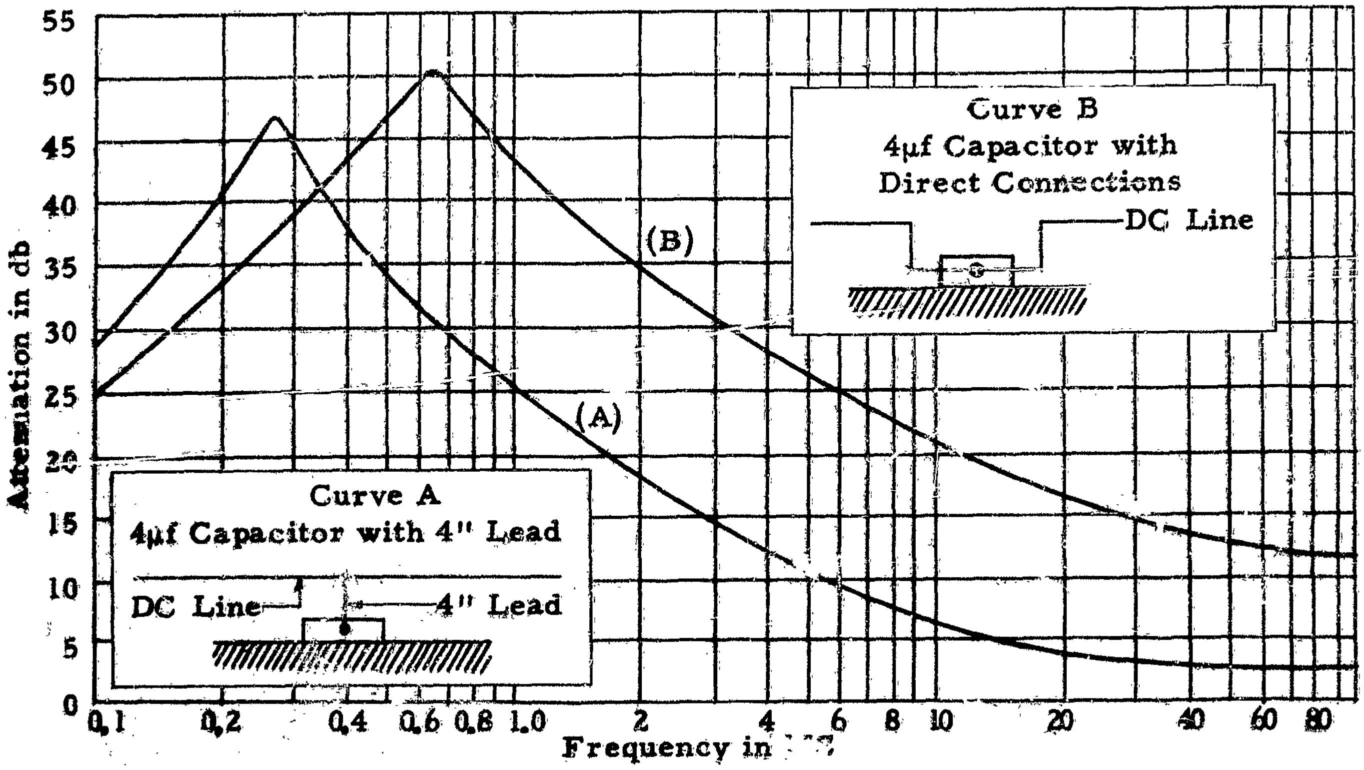


Fig. 3.1.1.1-A Difference in Capacitor Attenuation With Different Arrangement of Leads

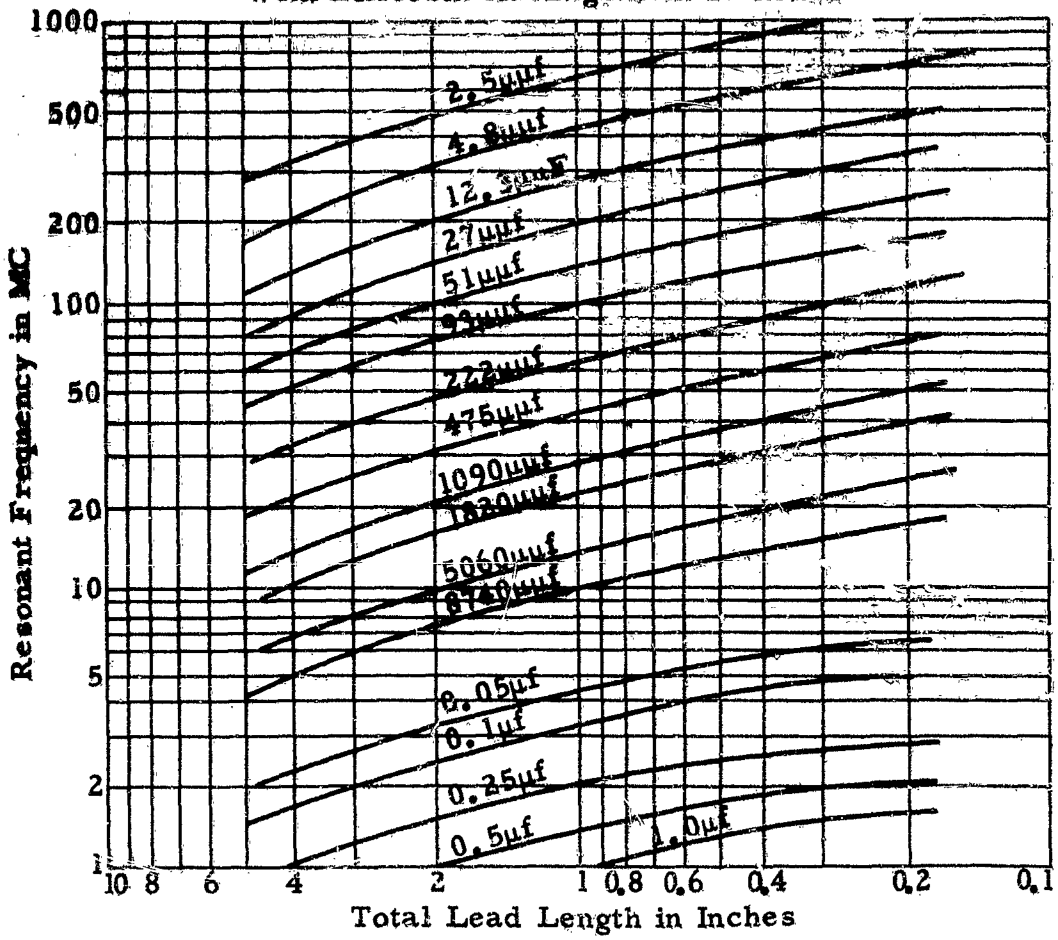


Fig. 3.1.1.1-B Series Resonance of Small Capacitors as a Function of Lead Length

When a capacitor is used with an alternating current generator, additional requirements must be met as follows:

(a) The resonance of the capacitor with the internal inductance of the generator at the fundamental or the lower harmonic frequencies must be avoided to prevent the build-up of dangerously high currents and voltages.

(b) The capacitance must not be so large as to increase the generator output sufficiently to cause overheating.

(c) The excitation of a generator with automatic voltage regulation must not be increased by the capacitor current more than the amount for which the regulator can compensate.

To meet these requirements, the maximum value of capacitance that can usually be used on alternating current generators is 0.5 µf with a value of 0.05 µf preferred for most applications. Since alternating current generators have no commutators, the interference generated by them is not as severe as that generated by direct current generators, and the reduction of radio interference afforded by a 0.05 µf capacitor is usually sufficient.

Another important factor in the installation of capacitors is provision for a good ground connection. Since the purpose of the capacitor is to provide a low impedance path, and since the grounding connection is in series with the capacitor itself, it is clear that the connection must be a low impedance bond at all frequencies at which the capacitor should be effective. A single-terminal capacitor in a grounded metal case is most effective. The grounding connection should be a bare metal to metal contact with high pressure maintained mechanically. An exception to this occurs when the negative lead of the capacitor must be connected directly to the negative side of the interference generator. In this case, a capacitor with two terminals should be selected, and both leads should be kept as short as possible.

Coupling between the lead carrying the interfering currents to the capacitor and the "clean" lead carrying only the desired currents must be avoided. In the case of a single-terminal capacitor, the two leads should be collinear and pointed in opposite directions. Under no circumstances should the two leads ever be bundled together or run parallel to each other.

Instead of using a single capacitor, a single inductor in series with the lead carrying interfering currents can also be used. This is very rarely done because usually more attenuation can be obtained with a shunt capacitor of comparable size and weight. The greatest disadvantage of the series inductor is that it must carry the full line current at the desired frequency. To obtain the amount of inductance necessary for adequate suppression of radio interference, an excessive amount of copper is usually required. Cores of magnetic material increase the inductance considerably, but if the desired frequency is 60, 400, or 800 cps, such cores introduce undesirable losses at the power frequency. For use in direct current leads, no such losses exist, but the direct current may produce saturation of the core, thus lowering the effective inductance.

Single coils have one advantage over single capacitances, which may, in

isolated instances, induce the designer to give them preference: Shunt capacitors, since they offer a low impedance to interfering currents, actually increase these currents in the generator, while series inductors, which offer a high impedance to the interfering currents, decrease these currents everywhere.

The design of inductors will be treated in detail in Paragraph 3.1.1.6. As far as the installation of inductance coils is concerned, the most important consideration is to keep the coils out of any interfering fields. If no interference-field-free location can be found, the coils must be shielded. If they are allowed to pick up any interference through inductive or capacitive coupling, their effectiveness is obviously lost.

The upper frequency limit of an inductance coil is determined by its distributed capacitance. At some frequency this capacitance will resonate with the inductance, and at frequencies above that, the effectiveness of the coil in suppressing radio interference decreases rapidly. The distributed capacitance of a 100-millihenry iron-core coil is about 3 to 15 μμf, depending on the kind of winding. The resonant frequency of this coil could be no higher than 250 kc, and it would not be suitable for radio interference suppression. An inductance of 100 μh is usually sufficient, and with proper care in design (see Paragraph 3.1.1.6) the resonant frequency of such a coil may be as high as 20 mc. For the suppression of frequencies above 20 mc, a single coil can be used only if inductance values below 100 μh are sufficient.

Finally, "brute-force filters" may be constructed by combining shunt capacitors and series inductances into L- or pi-sections as shown in Figure 3.1.1.1-C. These sections give higher attenuation than single elements and should be used when the attenuation afforded by single inductors or capacitors is insufficient.

The individual requirements for the installation of single capacitors or single inductors apply equally when such elements are used in combination. The requirement for good grounding connections acquire special significance for the pi-section, as is evident from Figure 3.1.1.1-D. Here a poor bond actually allows the interfering currents to by-pass the inductance coil and return to the line thus vitiating the purpose of the network.

The choice between a condenser-input and an inductance-input L-section is determined by the following considerations: The inductance-input L-section (Figure 3.1.1.1-C (1)) offers the higher input impedance to the interfering currents; therefore, it should be chosen whenever there is a reason for making it desirable to reduce the interfering currents before the point of application of the suppressing network. Such reasons might be the prevention of overloading of a generator, or the prevention of inductive coupling of these currents to other circuits. In the absence of such reasons, the condenser-input L-section (Figure 3.1.1.1-C (2)) should be chosen because it results in greater attenuation. This is proved for a somewhat special case, in Appendix IX.

In a pi-section, great care must be taken that there be no coupling, inductive or capacitive, between the two capacitors. Figure 3.1.1.1-E shows how such coupling allows the interfering currents to return to the line. Such coupling may be prevented by extreme care in the installation or, in severe cases, by complete shielding of at least one of the capacitors. More details about these procedures will be

found in the following paragraphs.

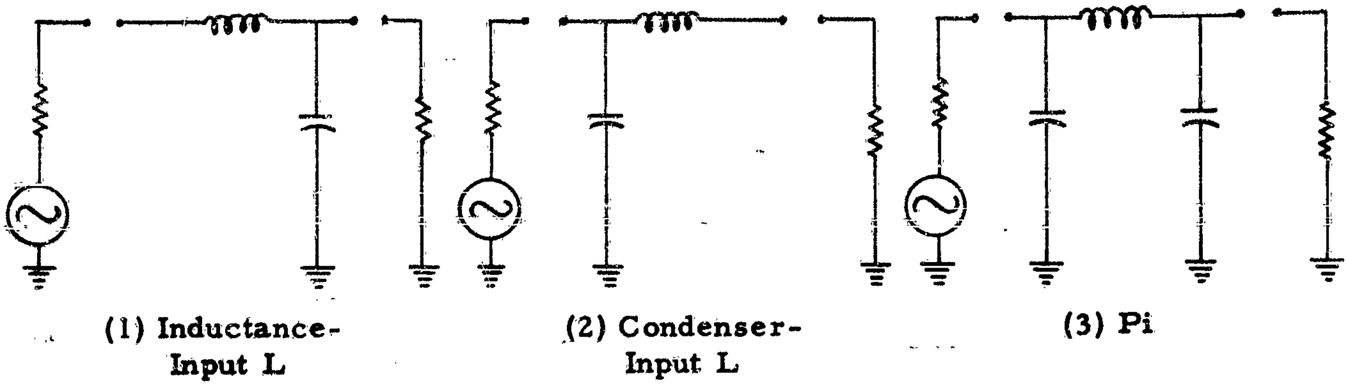


Fig. 3.1.1.1-C "Brute-Force Filter" Sections

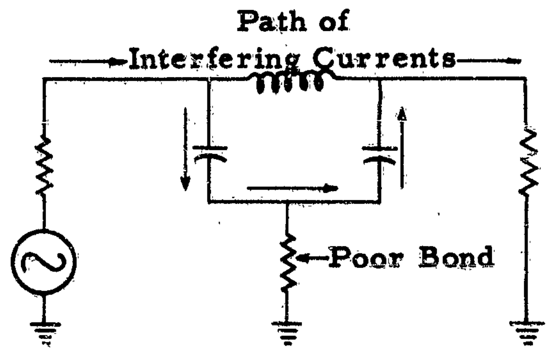


Fig. 3.1.1.1-D Effect of Poor Bond in Pi-Section

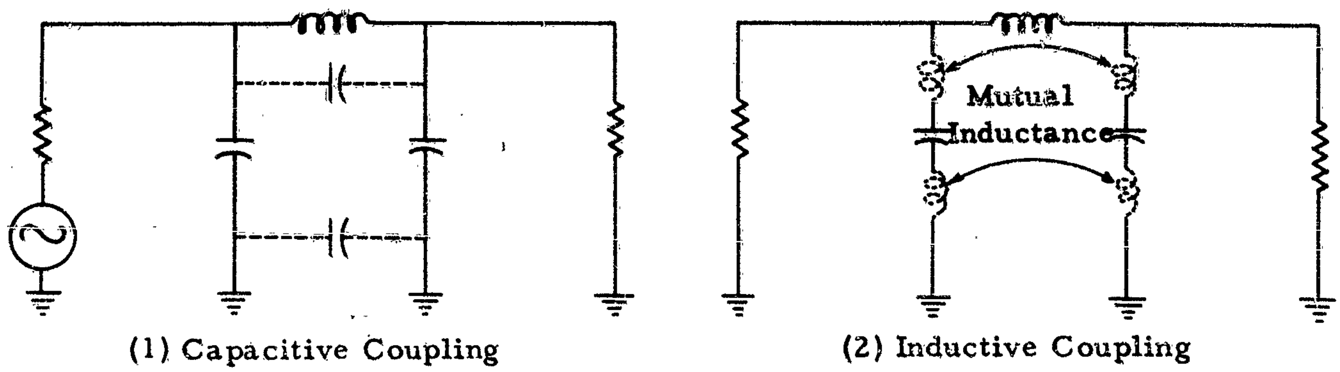
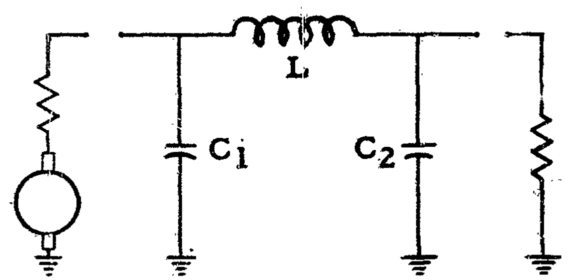


Fig. 3.1.1.1-E Coupling Between Condenser Leads in Pi-Section

Figure 3.1.1.1-F gives three designs of practical brute-force pi-sections. For complete suppression of all frequencies from 150 kc to 150 mc, the three filters must be used in series. These design instructions are approximate and should be used only as a guide.



Remarks

Determine wire size for coil from table on right.

Voltage rating of capacitors should be about twice generator or load voltage. (50 volts recommended for 28 volt systems).

	L	C_1	C_2
150 kc to 20 mc	12 Turns single cotton enameled magnet wire on 3/4" diameter form, close wound. Laminated iron core 3/8" x 5/8" with sufficient length to accommodate winding. Laminations 0.025" thick.	4 µf Single terminal (grounded metal case)	1.3 µf Single terminal (grounded metal case)
20 mc to 100 mc	10 Turns enameled magnet wire on 1/2" diameter form, space wound (equivalent to diameter of conductor).	0.5 µf Single terminal (grounded metal case)	0.01 µf mica (leads as short as possible)
100 mc to 150 mc	8 Turns enameled magnet wire on 1/4" diameter form, space wound (equivalent to diameter of conductor).	500 µµf mica (leads as short as possible)	500 µµf mica (leads as short as possible)

Max. Current in Amps	Wire Size	Diameter in Mils	k* No. of turns per inch for close windings: EN & SSC	k* No. of turns per inch for close windings: DSC & SCC
1.5	22	25.3	37	34
2	21	28.5	32.9	29.8
4	18	40.3	23.7	22.0
6	16	50.8	18.9	17.9
10	14	64.1	15	14.2
15	12	80.8	12	11.5
25	10	102	9.6	9.3
40	8	128	7.6	7.4
70	6	162	----	----
100	4	204	----	----
175	2	258	----	----

* Approximate only since thickness of insulation depends on type of insulation and manufacturer.

EN - Enamel
SSC - Single Silk Covered
DSC - Double Silk Covered
SCC - Single Cotton Covered

Fig. 3.1.1.1-F Design Instructions for Brute-Force Pi-Section in Direct Current Systems

3.1.1.2 POWER LINE FILTERS

Power line filters are used at the output of interference sources such as motors, generators, or inverters, and at the input power lines of receivers. Filters of this type are also required for the leads carrying power into shielded rooms. In this last case, the power frequency is usually 60 cps, but in all other cases the power frequency will be assumed to be 400 cps.

Power line filters for the suppression of radio interference are low-pass filters. If they are to operate efficiently, i.e., if they are to transmit the power frequency with little or no losses, a knowledge of the impedances between which they are to operate is essential for their design. An exact knowledge of these impedances is practically never available, but an approximate value can usually be obtained at the power frequency. Fortunately, as was explained in detail in Paragraph 1.8.2.3, the effect of an impedance mismatch in the pass band can be kept negligibly small by choosing the cut-off frequency high enough. As was pointed out there, to reduce the effect of the filter on the desired power currents to a minimum, the image impedance of the filter should be chosen near the geometric mean of the impedances between which it is to operate. In the attenuation band, an impedance mismatch usually introduces additional losses, a condition which is desirable, or at worst it introduces a reflection gain, which is never more than a small fraction of the total attenuation. Therefore, in the design of power line filters, it is only necessary to consider the impedances at the power frequency.

Most loads, such as receivers and lighting systems, may be assumed to have a purely resistive input impedance. This resistance may be computed either by dividing the normal input current into the nominal input voltage, or by dividing the normal input power into the square of the nominal input voltage:

$$R = \frac{V}{I} = \frac{V^2}{P} \tag{3-3}$$

depending on whether the load is rated in terms of current or power. For example, a 110-volt load rated at 500 ma would have a resistance of 110/.5 = 220 ohms. The same load might be rated at 55 watts, and its resistance would be computed as $110^2/55 = 220$ ohms.

Loads such as motors or relays usually have inductance as well as resistance. For practical purposes it is best to ignore the inductance and design the filter on basis of the resistance only. For this purpose, the exact value of the resistance may be obtained by dividing the square of the current into the power, $R = P/I^2$, but the error in using Equation (3-3) is small for the power factors usually encountered in aircraft equipment. Since an approximate value of resistance is sufficient, the errors caused by assuming unity power factor are not significant.

In conventional filter design, much attention is given to maintaining a good impedance match over the entire pass-band. The development of m-derived terminating half-sections was motivated entirely by this necessity. In power-line filters only one frequency is of importance in the pass-band. Therefore, wide-band matching is not a problem. However, it may still be desirable to use m-derived

sections in order to meet specific attenuation requirements.

The image impedance of a conventional low-pass filter of the constant-k or m-derived type is exactly equal to the design resistance R only at zero frequency. For other frequencies in the pass band, for a pi-section, it is given by:

$$Z = \frac{R}{\sqrt{1 - (f/f_c)^2}} \tag{3-4}$$

and for a T-section, it is

$$Z = R \sqrt{1 - (f/f_c)^2} \tag{3-5}$$

where Z is the input impedance, f the frequency, and f_c the cut-off frequency in the same units as f. For a radio interference filter, f_c may not be lower than 100 kc. With a power frequency, f, of 0.4 kc, it is obvious that for all practical purposes Z may be taken equal to R.

Once the cut-off frequency and the value of R have been decided upon, the design may proceed according to the procedure outlined in Appendix VII. For the installation of the filter, all considerations outlined in Paragraph 3.1.1.1 for the installation of capacitors, inductance coils, and "brute-force filter" sections apply. When the filter consists of more than one section, say two constant-k and one m-derived sections, coupling of any sort between sections must be prevented. Where large attenuation is required, each section should be enclosed separately in a complete shield.

Examples of the design of typical filters for use as radio interference filters are in Appendix VII.

3.1.1.3 HARMONIC SUPPRESSION FILTERS

Harmonic suppression filters are used at the output of transmitters to prevent any harmonic of the desired transmitter frequency from reaching the antenna. They are band-pass filters, or possibly low-pass filters, since the frequencies below the desired transmitter frequency are not usually of interest. The cut-off frequency of such a filter should be between the desired fundamental, f_d, and its second harmonic, $2f_d$. A value of 1.5 f_d is usually a good choice.

Harmonic suppression filters are inserted between the output of the transmitter and the transmission line feeding the antenna, or between the transmission line and the antenna. Assuming that the system is designed so that there would be a good impedance match without the filter at both the transmitter and the antenna ends, the filter should be designed to operate between impedances equal to the characteristic impedance of the transmission line. For coaxial lines, the commonly used impedances are 52 and 75 ohms.

Conventional filters using a combination of constant-k and m-derived sections are effective even at ultra-high frequencies. The only special consideration at the higher frequencies is that short sections of transmission lines must be used as filter elements.

The design objectives for a harmonic-suppression filter are the following:

(a) In the pass band the filter must perform electrically as though it were simply another section of conventional transmission-line cable (52 ohm or 75 ohm, or any other impedance which may be required).

(b) The attenuation characteristic should be such that a certain minimum must be exceeded over a very wide band. A good choice would be to require an attenuation of not less than 60 db between 10% above and four times the cut-off frequency (to attenuate the strong lower harmonics), and of not less than 30 db between four and ten times the cut-off frequency.

(c) In the pass band, the insertion loss should be less than (1/2) db and the voltage standing-wave ratio caused by the filter when properly terminated should be less than 1.10 over the complete pass-band range.

(d) The physical size and weight of the filter should be held to a minimum consistent with good engineering practice.

(e) The average and peak power capacity must be sufficient for the transmitter considered.

(f) The filter should be hermetically sealed to insure satisfactory operation under all atmospheric conditions encountered in flight.

If attenuations greater than those stipulated under Objective (b) are required, it is better to use two or more complete filters in series than to attempt the design of a filter meeting more severe requirements.

The individual filter elements are extremely critical and may be constructed so that they can be tuned after the complete filter has been assembled. One satisfactory method of accomplishing this consists of varying the capacity of the condensers by rotating the dielectric spacers between parallel plates. The spacers are shaped like cams so that their rotation effectively controls the dielectric constant and thus changes the capacitance. After tuning, all of the adjustments must be permanently locked.

It is found that in a filter of this type, the insertion loss in the pass band is due primarily to the reflection coefficient and not the losses due to resistive components. Therefore, the voltage standing-wave ratio in the pass band with the filter properly terminated is an excellent indication of the filter's pass-band performance and efficiency. To keep the reflection coefficient small over the entire pass band, the input impedance of the filter should be as constant as possible. Terminating m-derived half-sections can be constructed specially for this purpose. A value of m near 0.6 gives the best results. But if the image impedance of such a half-section is plotted as a function of frequency, it is seen that there is a rapid decrease in its

magnitude in the region from about 90 to 100 percent of the mathematical cut-off frequency. A portion of this region must necessarily be considered as belonging in the attenuation band, if low values of voltage standing-wave ratio are to be obtained in the pass band. Therefore, the design cut-off frequency, f_c', should be taken as a little lower than the mathematical cut-off frequency, f_c.

A good rule to follow is

$$f_c' = 0.965\, f_c \quad \text{or} \quad f_c = 1.035\, f_c' \tag{3-6}$$

To increase the sharpness of cut-off, a value of m a little below 0.6 for the terminating half-sections is desirable. If a value of 0.5 is chosen, it must be remembered that the image impedance of such a section is a little above the nominal image impedance, R_o, (i.e., the image impedance used in the design of the filter) for most of the pass band. A good choice for the relation between the actual load impedance, R_L, and R_o is

$$R_o = 0.925\, R_L \quad \text{or} \quad R_L = 1.081\, R_o \tag{3-7}$$

At frequencies above 70 or 80 mc, it becomes convenient to use short sections of coaxial transmission lines as filter elements. These lines are used to simulate the required values of lumped inductance and capacitance. The dimensions of the lines required may be computed in the following way:

Any uniform, lossless transmission line of characteristic impedance, Z_o, and electrical length, θ radians, can be exactly represented by the equivalent circuit shown in Figure 3.1.1.3-A.

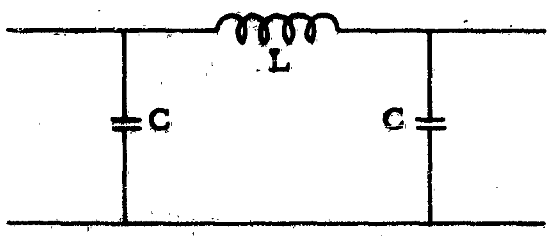


Fig. 3.1.1.3-A Equivalent Circuit of Transmission Line

The elements have the following values:

$$L = \frac{Z_o \sin\theta}{2\pi f} \tag{3-8}$$

$$C = \frac{\tan(\theta/2)}{Z_o\, 2\pi f} \tag{3-9}$$

where Z_o is in ohms and f, the frequency, in cycles per second. When θ is small (less than 15°), these values can be replaced to a good approximation by:

$$L \text{ (in } \mu\mu h) = 84.7 \; Z_o \, d \tag{3-10}$$

$$C \text{ (in } \mu\mu f) = \frac{42.3}{Z_o} \, d \tag{3-11}$$

where d is the actual length of the line in inches. It is seen that for an electrically short line, the equivalent lumped parameters are independent of frequency, which is an important requirement if the line is to simulate constant elements. It is necessary, therefore, that the line be less than 15-electrical degrees long at all frequencies for which the filter is to maintain its essential properties.

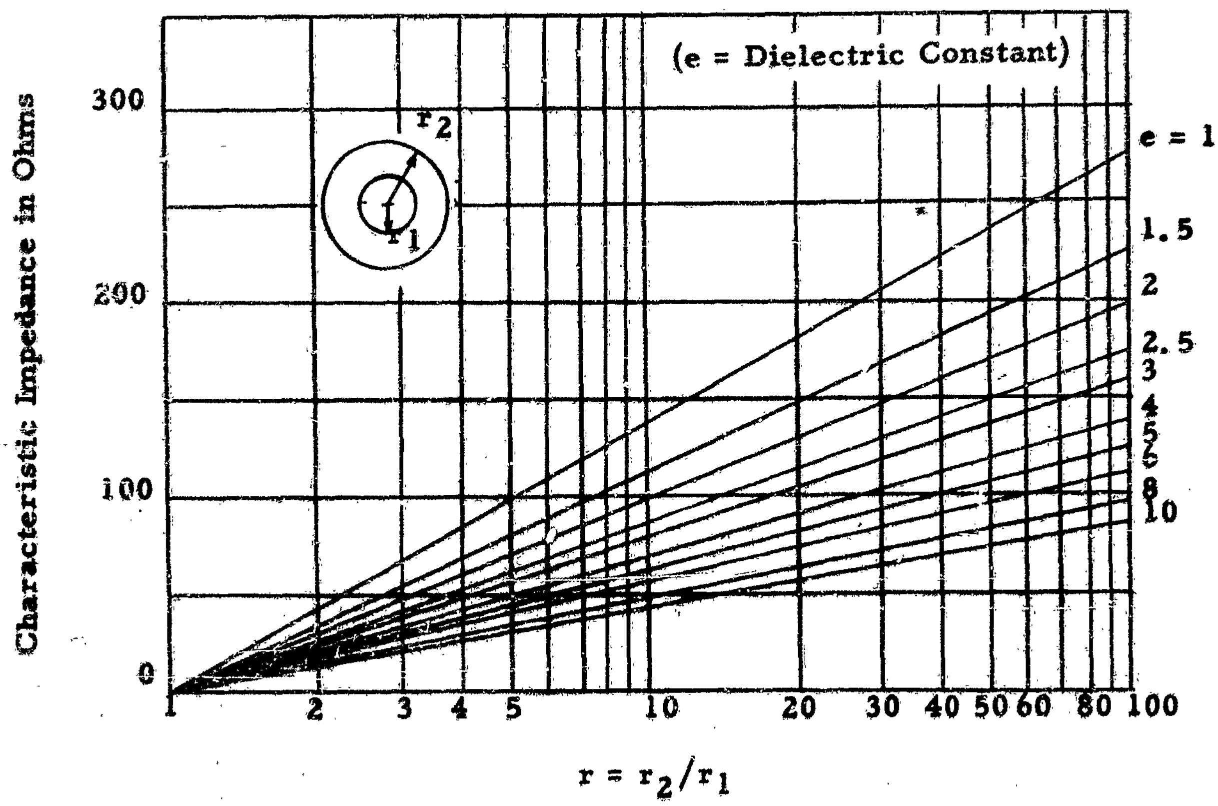


Fig. 3.1.1.3-B Characteristic Impedance of Coaxial Transmission Lines

The characteristic impedance, Z_o, of a coaxial transmission line using a solid dielectric is plotted in Figure 3.1.1.3-B as a funct n of the ratio r of the inner diameter of the outer conductor to the outer diameter of the inner conductor for several values of the dielectric constant e of the medium between the two conductors. The analytical expression for the characteristic impedance is

$$Z_o = \frac{138}{\sqrt{e}} \log_{10} r \tag{3-12}$$

An abrupt change in either the inner or outer conductor diameter of a coaxial transmission line causes a distortion of the field which produces an admittance that can be represented electrically as a shunt capacitance across the line at the place of discontinuity. The value of this discontinuity capacitance is appreciable and must be considered in high frequency filter design where short sections of transmission

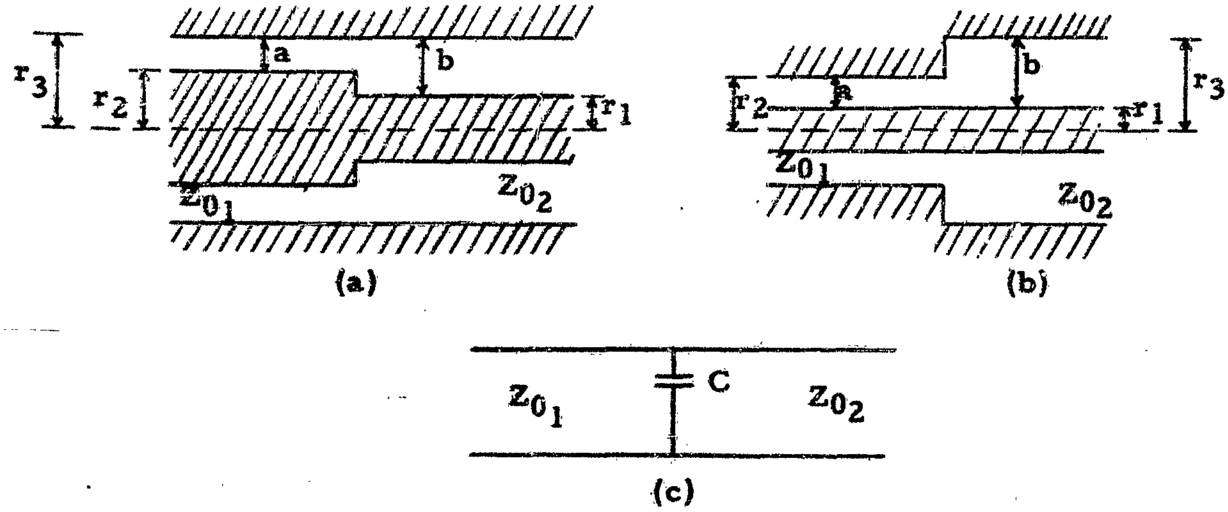


Fig. 3.1.1.3-C Discontinuities in Coaxial Lines and Equivalent Circuit

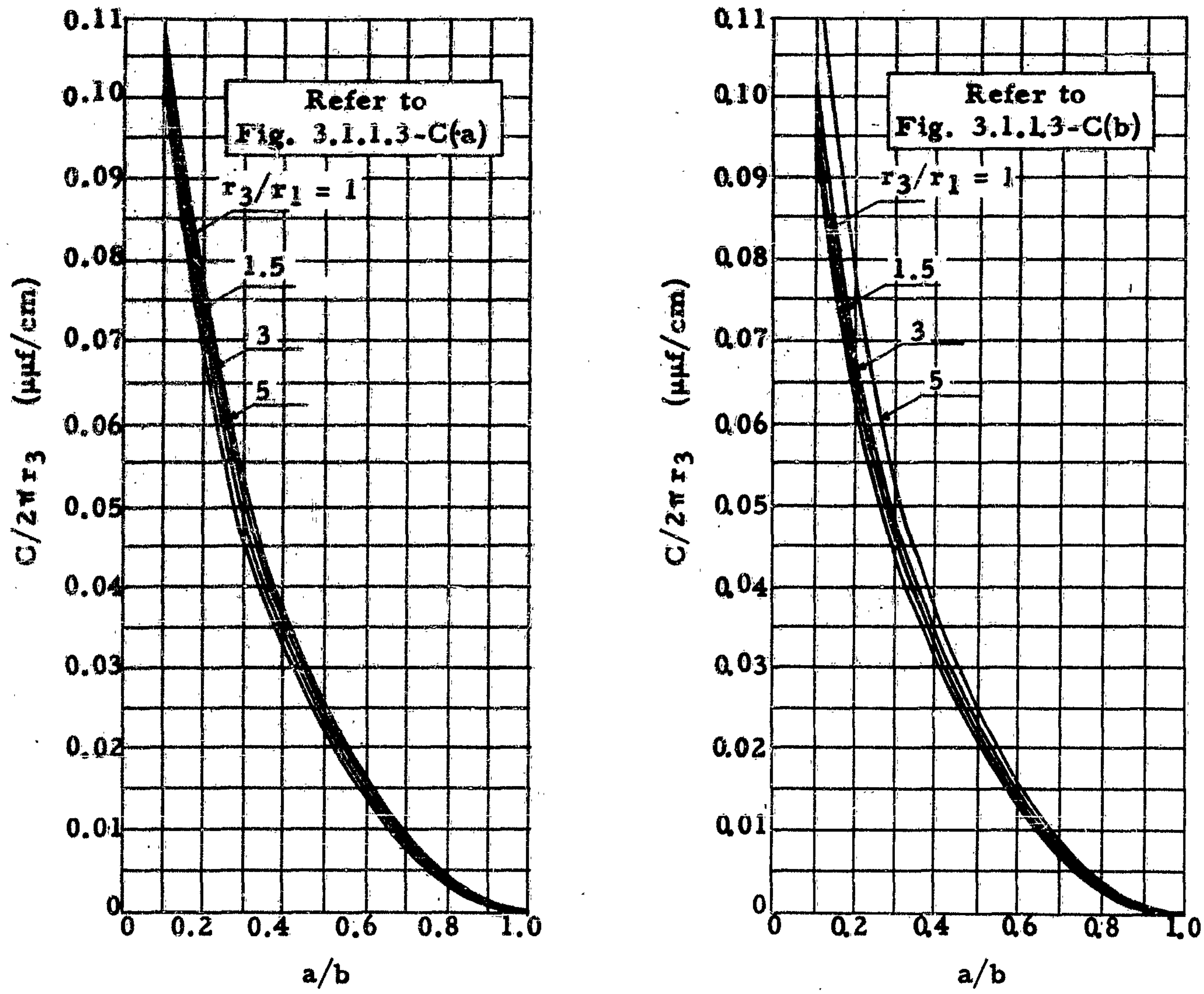


Fig. 3.1.1.3-D Curves for Computation of Discontinuity Shunt Capacitance in Coaxial Lines

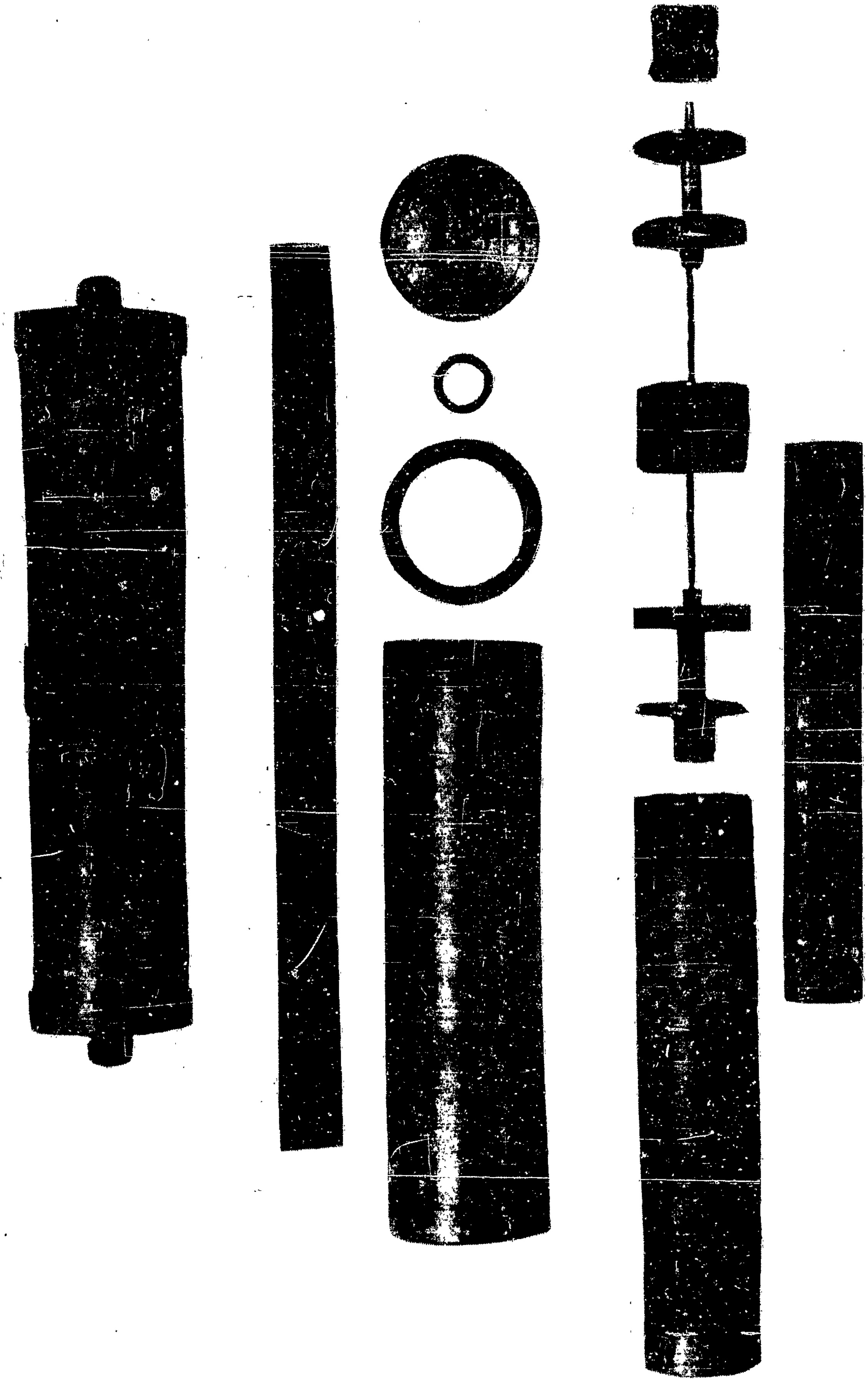


Fig. 3.1.1.3-E Exploded View of a Typical Low Pass Radio Frequency Filter
(52 ohm image impedance, 500 mc cutoff frequency,
50 db attenuation at frequencies over 550 mc)

line are employed as lumped elements. Two such discontinuities are pictured in Figure 3.1.1.3-C together with their equivalent circuit. The values of the shunt capacitances may be found from the curves on Figure 3.1.1.3-D. In each case, to obtain the actual shunt capacitance in micromicrofarads, the value read from the curve must be multiplied by the circumference of the unchanged conductor measured in centimeters. If there is a simultaneous discontinuity in both the outer and the inner conductor, an approximate but very good value of the equivalent shunt capacitance is obtainable by breaking it into two parts and using the same curves as before. That is, the capacitance is the sum of the two capacitances obtained by assuming first that the inner conductor remains continuous while the outer one changes, and then that the outer conductor remains continuous while the inner one changes.

An exploded view of a filter designed and constructed in accordance with the procedure outlined in this paragraph is shown in Figure 3.1.1.3-E.

Low-pass filters of the type described here are also sometimes useful in the antenna input circuit of receivers. While in a well-designed receiver the antenna input circuit acts as a band-pass filter attenuating all frequencies above and below its pass band, it may happen that a very strong signal outside the pass band overrides the desired signal despite this attenuation, so that an additional external filter becomes necessary.

3.1.1.4 DESIGN OF FILTER CAPACITORS

Capacitors for use in radio interference suppression networks must meet the following mechanical requirements:

(a) Extreme ruggedness. Must be unaffected by severe shock.

(b) High resistance to extreme altitudes, temperatures, and humidities.

(c) Durability. Must retain their mechanical and electrical properties without maintenance indefinitely.

(d) Ability to resist corrosion.

The electrical requirements are that they have the specified capacitance with minimum weight and space, that they be capable of withstanding the operating voltages without danger of flash-over, and that they neither produce nor be affected by external electric or magnetic fields. They must also preserve their electrical properties over as wide a frequency range as possible.

These requirements make it mandatory that the capacitors be hermetically sealed, if at all possible, unless they are part of a unit which is hermetically sealed as a whole. In general, the design should follow the requirements of the joint Army-Navy specification of date of issue at the time the design is formulated.

If hermetical sealing is impossible, some sealing compound must be used to protect the capacitors against the effects of high temperatures and humidities. Dow Corning 4 compound has been widely used for this purpose. Recent tests by the U. S. Navy have shown that, due to its "flow" properties and loss of insulation resistance

with absorption of moisture, this compound cannot be relied upon as a means of sealing an open-type, oil-dielectric capacitor. Breakdown and failure occurred frequently in these capacitors because the Dow Corning 4 compound mixed with the oil dielectric resulting in deterioration of the capacitor. Dow Corning 4 compound is satisfactory when used with synthetic rubber, neoprene, or phenolic insulating materials.

Capacitors to be used in suppression networks should always be of the non-inductive type. For example, a rolled impregnated-paper capacitor should make use of the extended foil construction as shown in Figure 3.1.1.4.

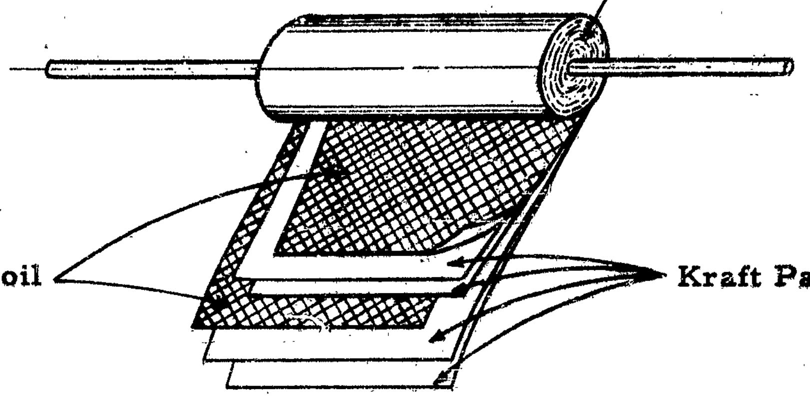


Fig. 3.1.1.4 Construction of Non-Inductive Capacitor

Capacitors used as shunts to ground should be of the single-terminal type with the metal case grounded. An exception to this occurs for low-capacitance mica condensers completely embedded in a plastic mold. Since the mold is not conducting, it cannot be grounded. Provision should be made to mount the capacitor at the point where the lead leaves the mold.

Capacitors to be used as series elements should be constructed so that the two leads leave the case or mold at opposite sides and so that connections can be made with a minimum length of leads. The internal lead length must also be kept at a minimum in order to keep the series inductance as small as possible.

3.1.1.5 FEED-THROUGH CAPACITORS

The "feed-through" capacitor differs from the conventional, or "lead-type", capacitor in that all ground leads have been completely eliminated. As shown in Figure 3.1.1.5-A, it consists of a feed-through bus that passes through the center of the capacitor section which is rolled in the extended foil manner. Alternate foils on each side of the feed-through bus are soldered together; one set is soldered to the housing and the other set is soldered to the feed-through bus. This type of a configuration limits the inherent inductance from line to ground to an almost negligible value thus raising the resonant frequency of the capacitor and extending its useful frequency range of effective by-passing.

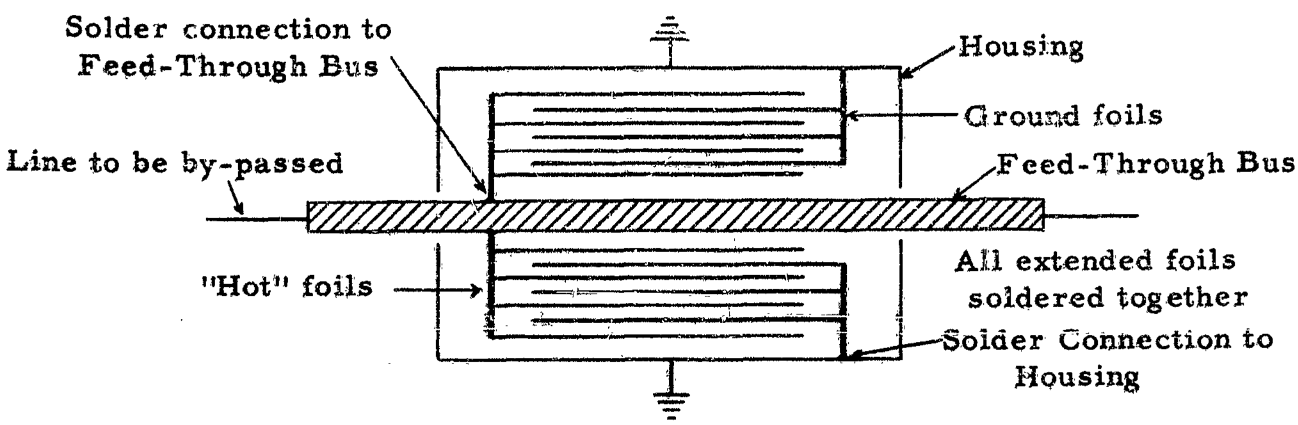
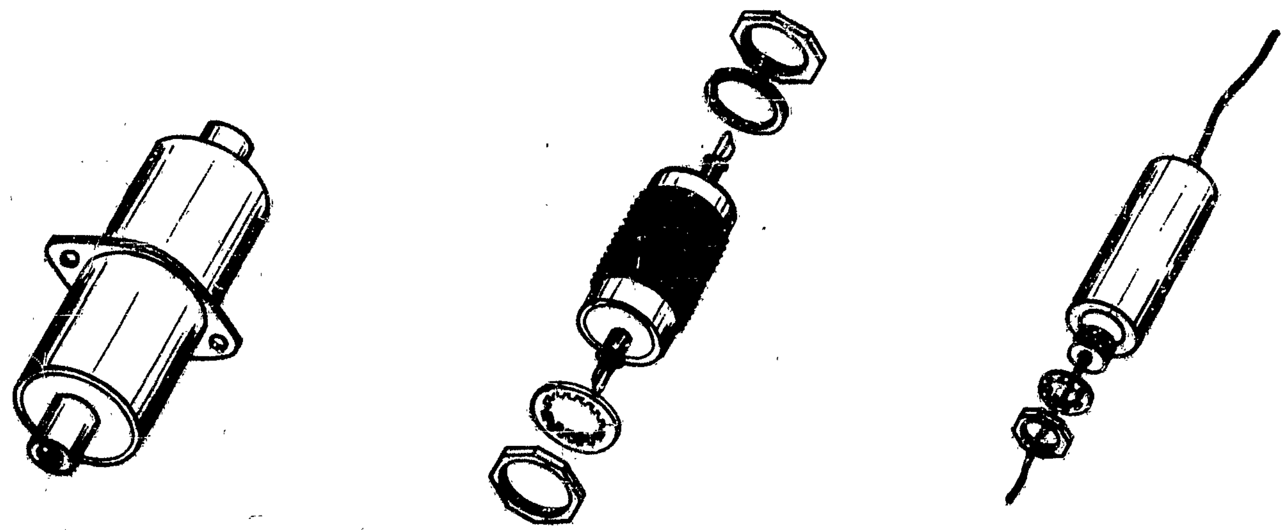


Fig. 3.1.1.5-A Schematic Diagram of a Feed-Through Capacitor



(1) Mounting Flange (2) Threaded Surface (3) Threaded Stud

Fig. 3.1.1.5-B Methods of Mounting Feed-Through Capacitors

The feed-through capacitor is essentially a three-terminal element since the line from which the radio frequency currents are to be by-passed must be broken and the capacitor inserted between the separated ends. The feed-through capacitor functions most effectively when mounted through a shield wall so that contact to ground is afforded along a continuous symmetrical line around the circumference of the housing. This minimizes the inductance and resistance from the housing to ground. Furthermore, the shield isolates the input and output leads of the capacitor from each other. Because of this, feed-through capacitors find their greatest usefulness at the exit points of leads from shielded enclosures containing interference sources. In this application they are far superior to any other single element though their insertion loss is not as great as that of a well-designed filter.

One method of mounting, as shown in Figure 3.1.1.5-B (1), makes use of a flange, an integral part of the capacitor wall, which is screwed to the shield. Better results are obtained, however, by using "press-in" or "cone" capacitors. The press-in capacitor is cylindrical in shape having an excess diameter of about a thousandth of an inch, while the cone capacitor, as the name implies, is conical. Both of these

capacitors are forced into a bored opening under great pressure. Another method of mounting, consistent with good results, employs a condenser cylindrical in shape whose surface has been threaded as shown in Figure 3.1.1.5-B (2). This capacitor is inserted into an opening in the shield and is held securely to both sides of the shield by a washer and nut arrangement. A similar type of mounting, shown in Figure 3.1.1.5-B (3), employs a smooth cylindrical condenser with a threaded mounting stud which is inserted into an opening in the shield and held securely to the shield by a lock washer and nut.

An analysis of the insertion loss - frequency curves of the feed-through, lead-type, and the ideal capacitor (i.e., one without any resistance or inductance) as shown in Figure 3.1.1.5-C, clearly shows the superiority of the feed-through capacitor. The dip below the ideal observed in the curve of the feed-through capacitor is typical of all capacitors of this type and usually occurs in the 50-600 megacycle range. The cause of the dip is not precisely known, but it is probably caused by a complex interaction of currents and fields in and surrounding the different layers of foil. Despite the dip, the feed-through capacitor is superior to the lead-type capacitor since no resonant effects are apparent and the insertion loss continues rising after the dip.

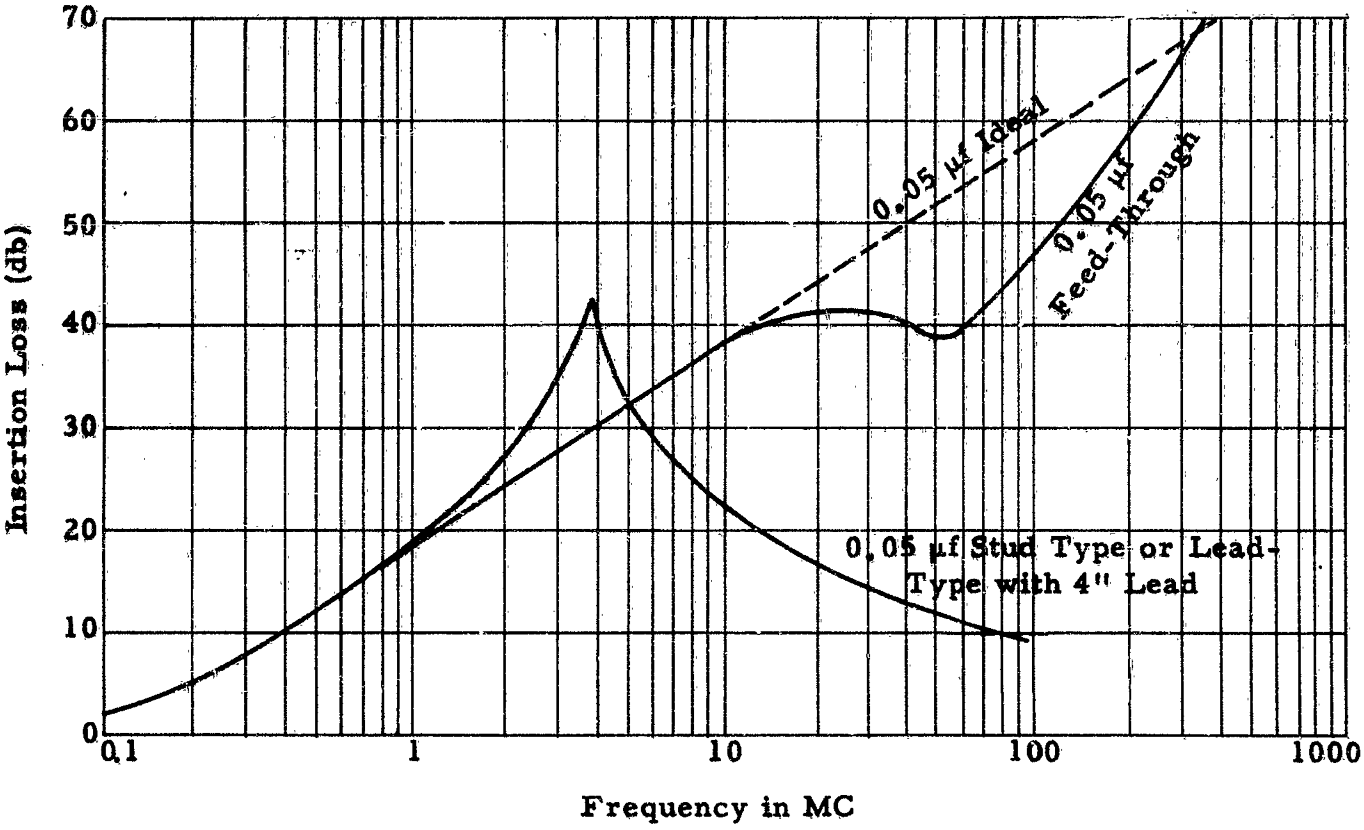


Fig. 3.1.1.5-C Curves of Insertion Loss as a Function of Frequency Comparing Feed-Through and Lead-Type Capacitors with Ideal Capacitors

Production samples of feed-through capacitors show wide variations in their insertion loss - frequency curves in the region approaching 1000 megacycles. This variation is probably caused by the radio-frequency resistance which at these frequencies is equal to or greater than the capacitive reactance of the capacitor thereby

exerting the greater influence on the shape of the curve.

Another type of a feed-through capacitor, shown in Figure 3.1.1.5-D, consists essentially of a feed-through bus passing through the center of the capacitive section which consists of discs rather than rolled foil. Alternate discs on both sides of the feed-through bus are connected to the housing while the remaining discs are connected to the feed-through bus. This type of configuration is best adapted for use with ceramic dielectric discs whose surfaces have been coated with silver to form the conducting plates. Large capacitors of this type have not as yet been satisfactorily produced. However, experimental models with values greater than 0.1 microfarads have been developed.

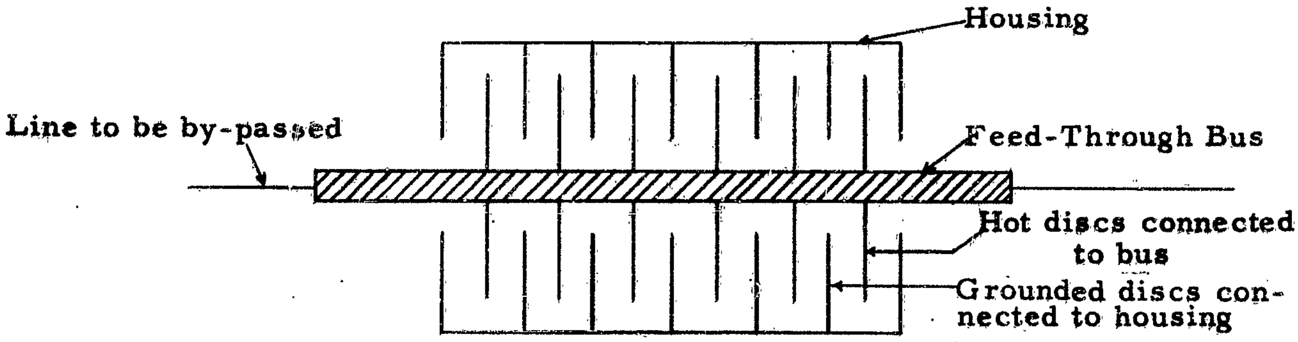


Fig. 3.1.1.5-D Schematic Diagram of a Stacked-Discs Type of Feed-Through Capacitor

3.1.1.6 DESIGN OF FILTER INDUCTORS

Inductors used in radio interference suppression networks must meet the same mechanical requirements as capacitors, which are enumerated in Paragraph 3.1.1.4. The electrical requirements are that they have the specified inductance with a minimum weight and space, that they be capable of passing the operating current without excessive heating, and that they neither produce nor be affected by external electric or magnetic fields. They must also preserve their electrical properties over as wide a frequency range as possible.

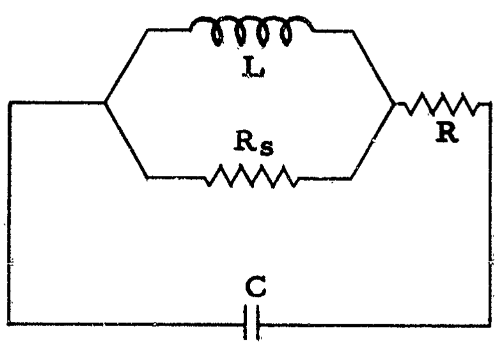


Fig. 3.1.1.6-A Equivalent Circuit of Inductance Coil

Figure 3.1.1.6-A shows the equivalent circuit of an inductance coil. The quantity L is the inductance, R_S represents the shunting effect of the losses in the surrounding medium, R is the actual winding resistance taking into account the skin

effect, and C is the distributed capacitance. The distributed capacitance may have losses, but these are usually negligible compared to those already mentioned. For direct and low frequency alternating currents, the core losses are very small and R approaches the direct current value of the winding. For use in an effective suppression network, C should be as small as possible so that the anti-resonant frequency, above which the inductor behaves like a capacitance and will lose its effectiveness, is as high as possible. The inductance L should be large for the frequencies of the interfering currents, but low for the desired power frequencies. The value of R is not important as far as the undesired frequencies are concerned, but a low value of R is desirable in order to keep the voltage drop across R as small as possible for the desired currents.

At the radio frequencies which are to be suppressed, practically all the losses associated with the inductance coil occur in R_s. Thus the "Q" of the coil is determined primarily by R_s. It is known from filter analysis that the effect of dissipative elements is negligible in all cases except in the vicinity of the critical frequencies (the cut-off frequencies and the frequencies of infinite attenuation). Since radio interference suppressing filters or networks usually operate well beyond cut-off, relatively low values of R_s may be tolerated. Values of "Q" as low as three reduce the effective impedance of a coil by only 30 percent as compared to its value for the ideal "Q" of infinity.

It is very important to reduce the capacitance C as much as possible, and this may be done by connecting several coils in series so that the various capacitances of the individual coils are in series. Since this also lessens the inductance of the coil, because it eliminates some of the mutual inductance between the turns of different coils, a compromise must be found.

Multiple-layer coils should be avoided since their capacitance is appreciably greater than that of single-layer coils. The size of wire to be used depends on the current to be carried. The wire size should be chosen so that 100 percent overload can safely be carried except in special cases where circumstances may require a larger factor of safety. In determining the current, the radio interference currents may be neglected in comparison with the power currents. The correct wire size may be found in the table of Figure 3.1.1.1-F, page 3 - 10.

For inductances below about 1000 microhenries, usually no magnetic material need be used unless special space requirements exist. The inductance of a coil is a very complicated function of its physical dimensions and the number and arrangement of the turns. However, the following equation may be relied on to give correct results for a single-layer air-core coil within about 2 percent provided that the ratio of diameter to length, a/b, is neither very large nor very small as compared to unity.

$$L = \frac{0.5\, a^2 b^2 k^2}{9a + 20b} \tag{3-13}$$

The inductance, L, is in microhenries, the average coil diameter, a, is in inches, (see Figure 3.1.1.6-B), the length of winding, b, is in inches (see Figure 3.1.1.6-B), and k is the number of turns per inch. Figure 3.1.1.6-C is a nomograph based on Equation (3-13) for the rapid computation of either the inductance of

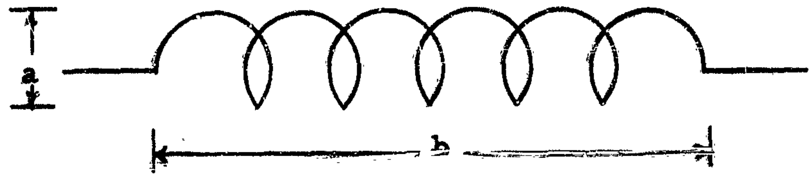


Fig. 3.1.1.6-B Dimensions of Single-Layer Air-Core Coil

a coil of given dimensions, or the dimensions of a coil for a given inductance. The use of this nomograph is as follows:

To find the inductance of a coil whose dimensions and number of turns are known, find the reference point on the "turning" scale by drawing a straight line through the approximate points on the "b/a" and the "a" scales. Then draw a straight line through this reference point and the approximate point on the "k" scale and extend this line until it intersects the "L" scale.

The following example illustrates the use of the nomograph in solving the above type of problem. Refer to Figure 3.1.1.6-C. Find the inductance of a coil close wound with 13 turns of No. 22 enamel wire (37 turns per inch), if its diameter and length are 3.5 and 0.35 inches, respectively.

Connect b/a = 0.1 and a = 3.5 with a straight line and rotate this line about the reference point on the turning scale to k = 37, and read L = 30 microhenries.

The length and the number of turns of a coil of known diameter, inductance, and current carrying capacity can be found by the use of the nomograph by finding first the number of turns per inch from the table of Figure 3.1.1.1-F, page 3 - 10. Then, a straight line is drawn between the appropriate points on the "k" and "L" scales to locate the reference point on the "turning" scale. This point, in turn, is connected to the appropriate point on the "A" scale by a straight line which is extended until it intersects the "b/a" scale. Since the value of "A" is known, the value of "b" can be readily calculated and used to obtain the total number of turns, N, by means of the relationship, N = kb.

The following example illustrates the use of the nomograph for the above conditions. Refer to Figure 3.1.1.6-C. Find the length and the number of turns of a 30-microhenry coil, designed to carry a maximum current of 1.5 amperes, and whose diameter is 3.5 inches. Figure 3.1.1.1-F shows that 37 turns per inch of No. 22 enamel wire is required.

Connect k = 37 and L = 30 with a straight line and rotate this line about the reference point on the turning scale to a = 3.5 and read b/a = 0.1. Since a = 3.5 inches, b = 0.35 inches. Therefore, the number of turns required is 0.35 x 37 = 13 turns.

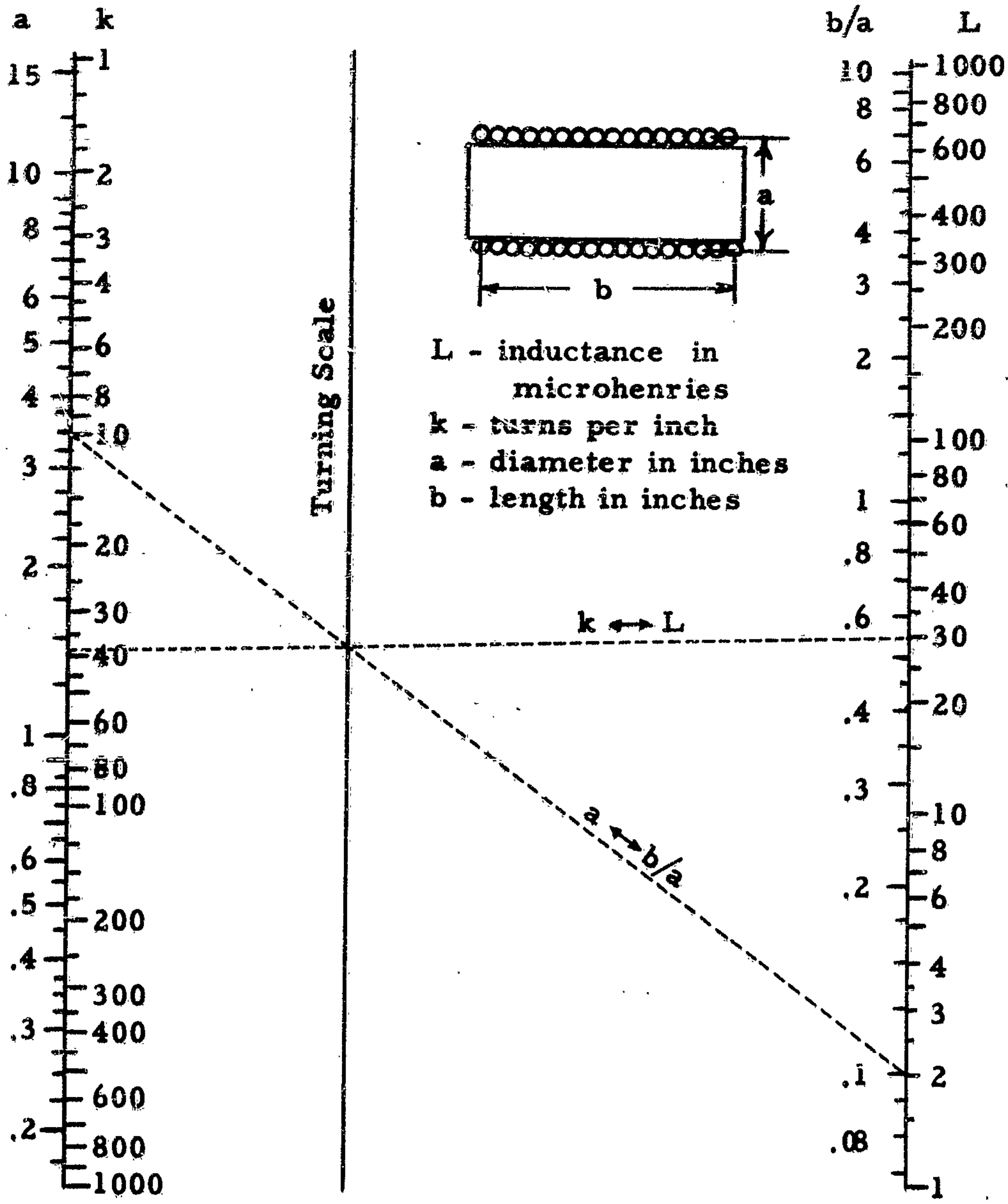


Fig. 3.1.1.6-C Nomograph for Single-Layer Coil Design

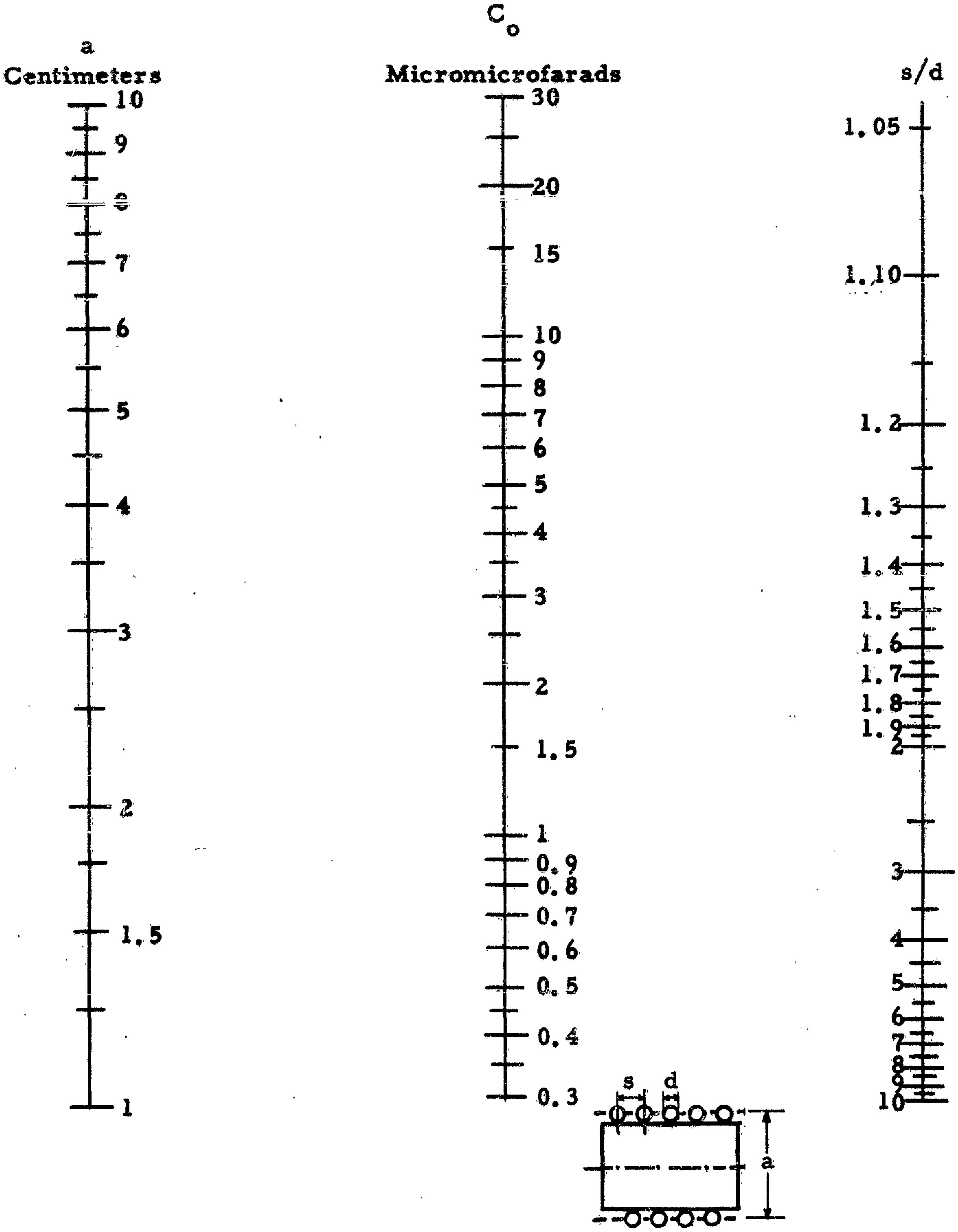


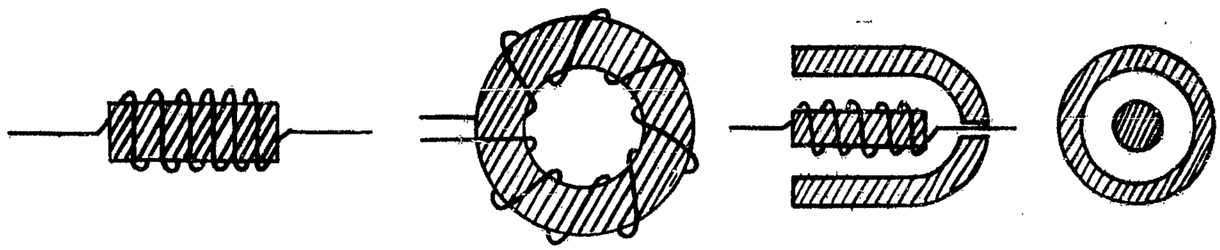
Fig. 3.1.1.6-D Nomograph for Determining Distributed Capacitance of Single-Layer Coils

If space is ample, it is usually desirable to make "k" smaller than its maximum value read from the table of Figure 3.1.1.1-F because a decrease of "k" will reduces the distributed capacitance of the coil. This reduction of the value of "k" is obtained by space winding rather than close winding the coil. It should be noted, however, that an error in the value of the inductance is introduced because the value of "k" used in the nomograph of Figure 3.1.1.6-C is based on a close-wound coil.

The distributed capacity, C_o, of a single-layer coil, shown in cross-section in Figure 3.1.1.6-D, can be readily found by the use of the nomograph of Figure 3.1.1.6-D, if the diameter of the coil "a", the diameter of the wire used "d", and the distance between the centers of adjacent turns "s" are known. The procedure for the use of this nomograph is as follows:

Draw a straight line connecting the appropriate points on the "s/d" and "a" scales. The point of intersection of this line and the "C_o" scale gives the value of the distributed capacity. If the diameter of the coil is "m" times larger or smaller than the values on the "a" scale, the value of the distributed capacity is also "m" times larger or smaller than the value read on the "C_o" scale. In this case, "m" is any constant factor.

If more inductance is needed in a given space than can be obtained in an air-core coil, a core of magnetic material must be used. Suitable coils using magnetic cores may take many forms. Several important arrangements are shown in Figure 3.1.1.6-E. Arrangement (1) in this figure shows a simple straight solenoid whose greatest merit is ease of manufacture. Arrangement (2) shows a toroid, which provides a satisfactory filter inductor as the external flux is almost nil. The distributed capacitance may be satisfactorily low; and, for a given space requirement, it provides a good value of inductance. Finally, arrangement (3) is an improved design which makes excellent use of available space, allows a very close-fitting shield to be used, and has a relatively low value of distributed capacitance.



(1) Straight (2) Toroid (3) Cup Core

Fig. 3.1.1.6-E Designs of Magnetic Core Inductors

If the power current is direct current, the saturation effects of the load currents must be carefully checked. In arrangement (1), the large airgap, plus the intrinsic airgaps due to the interstices between iron particles in the case of powdered cores, generally prevents saturation. A toroidal core of laminated strips must have an airgap to prevent saturation. A toroidal core of powdered iron may be free from saturation effects for moderate load currents, but may still require an airgap for

very large load currents. In arrangement (3), special care must be taken, since saturation may easily occur. This is particularly important should end-caps of magnetic material be added.

A magnetic slug in a straight solenoid will increase its inductance by a factor of about two to four. The inductance of a toroid of rectangular cross-section with or without intrinsic airgap, but without external airgap, may be found from the equation,

$$L = 0.00508\, N^2\, b\, \mu_d\, \ln (r_2/r_1) \qquad (3\text{-}14)$$

where L is the inductance in microhenries, b is the core width in inches, μ_d is the average incremental permeability, r_1 and r_2 are the inside and outside radii, respectively, and N is the total number of turns. See Figure 3.1.1.6-F. The incremental permeability must be used because the quantity of interest is the incremental inductance for small alternating currents superimposed on a large direct current. Curves of μ_d as a function of the magnetization, H, in ampere-turns per inch, are plotted in Figure 3.1.1.6-G. The quantity H may be computed from the equation,

$$H = \frac{NI}{2\pi r_{av}} = \frac{NI}{\pi(r_1 + r_2)} \qquad (3\text{-}15)$$

where I is the direct current in amperes.

For toroids with round cross-section, the inductance is

$$L = 0.0319\, N^2\, \mu_d \left(r_m - \sqrt{r_m^2 - \frac{b^2}{4}} \right) \qquad (3\text{-}16)$$

where r_m is the mean radius of the toroid and b is the radius of the cross-section, both in inches, as shown in Figure 3.1.1.6-H.

Another design, which is particularly useful for frequencies above about 50 mc, is shown in Figure 3.1.1.6-I. It consists of a straight conductor surrounded by an annular core. It is essentially equivalent to a one-turn toroid, and Equations (3-14) and (3-15) may be used with N taken as unity. The package-ability, low voltage drop for direct current, and extremely low capacitance combine to make this an ideal design when a very large inductance is not required. The optimum inductance is obtained by using several layers of different material in the annular core. The design of such an inductor may be carried out with the aid of the curves of Figure 3.1.1.6-J.

It is assumed that the outside diameter of the core is known from a consideration of the available space. The direct current determines the wire size and thus the inside diameter of the core. The range of values of H may then be determined from the curves of Figure 3.1.1.6-J, which give the magnetic field intensity at any arbitrary distance, r, from the center. The materials having the largest μ_d may be chosen from the curves of Figure 3.1.1.6-G. Note that the intersection of two curves on this graph marks the value of H for which one material becomes better

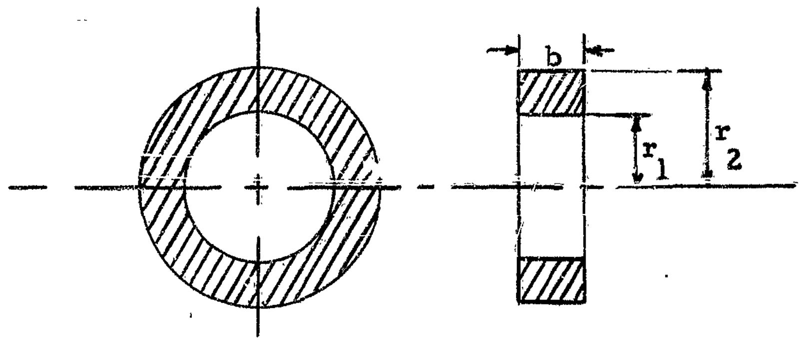


Fig. 3.1.1.6-F Dimensions of Toroid of Rectangular Cross-Section

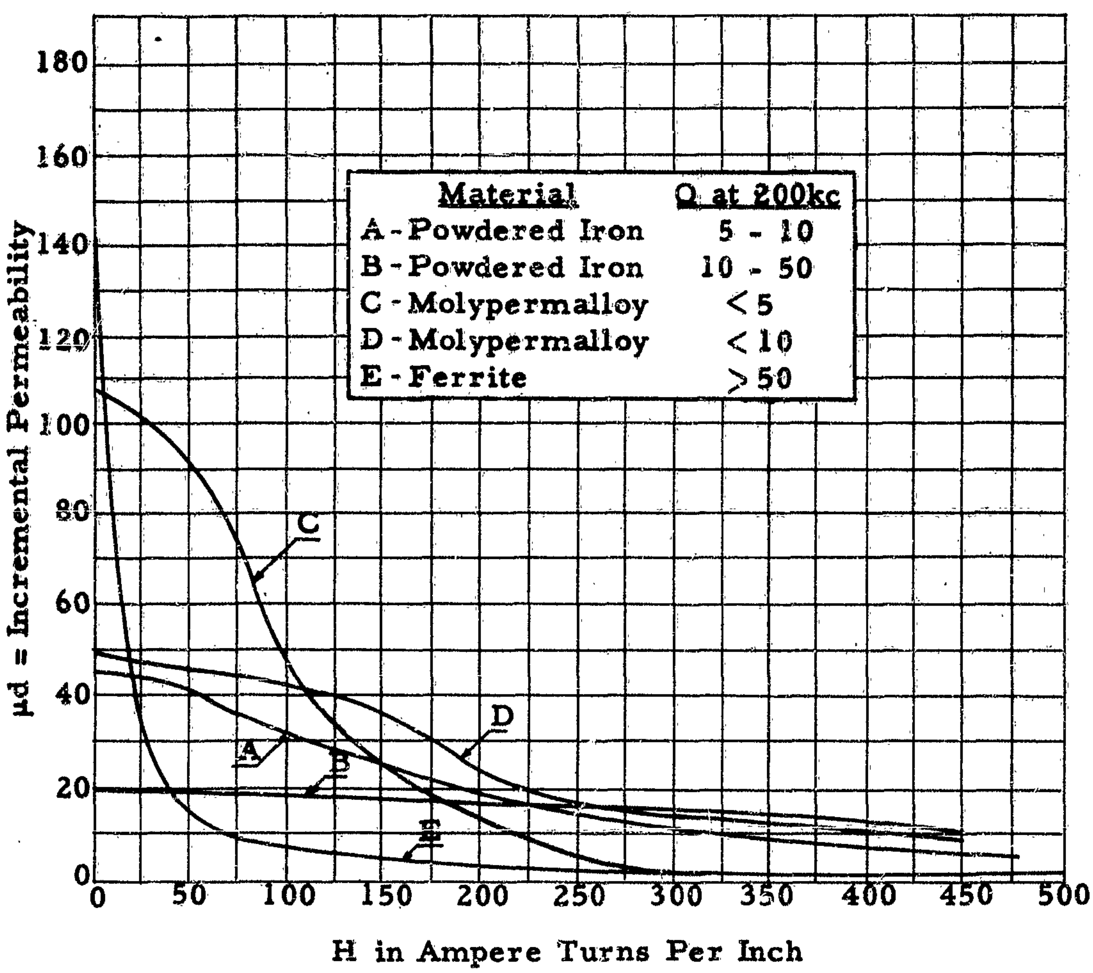


Fig. 3.1.1.6-G Incremental Permeability as a Function of Magnetic Field Intensity for Different Materials

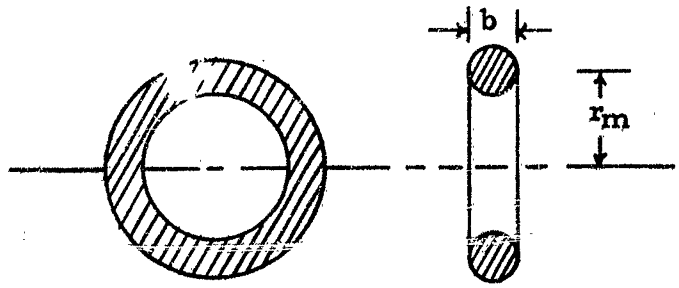


Fig. 3.1.1.6-H Dimensions of Toroid of Round Cross-Section

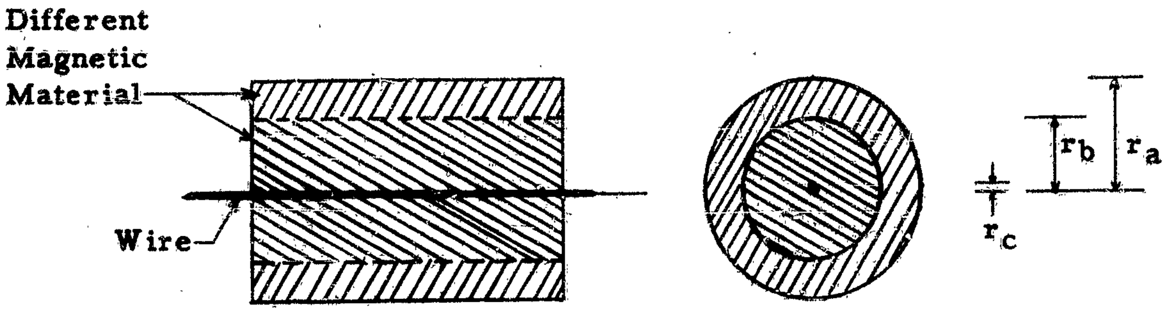


Fig. 3.1.1.6-I High Frequency Inductor

than another. Thus, Figure 3.1.1.6-J may be used to determine the radius or radii at which one material should be replaced by another. Finally, the inductance per unit length is found by adding the inductance values for the individual cores:

$$L' = 0.00508 \left(\mu_{d_1} \ln \frac{r_a}{r_b} + \mu_{d_2} \ln \frac{r_b}{r_c} + \ldots \right) \qquad (3\text{-}17)$$

where L' is in microhenries per inch and r_a, r_b, and r_c are explained in Figure 3.1.1.6-I.

As an example, let it be required to determine the optimum inductance per unit length of a conductor to carry 10 amperes direct current, with a maximum diameter of 0.6 inches available for the core. From the table in Figure 3.1.1.1-F, a wire size No. 14 is chosen. Allowing a hole of 0.08 inches in diameter, the radius of the core varies from 0.04 to 0.3 inches. From Figure 3.1.1.6-J, this corresponds to a variation in the field intensity, H, from 40 to 5.5 ampere-turns per inch. Figure 3.1.1.6-G shows that, for H having values below 12 ampere-turns per inch, material E (ferrite) has the highest incremental permeability, while, for H between 12 and 40, material C (molypermalloy) should be chosen. The value H = 12, from Figure 3.1.1.6-J, corresponds to a radius of 0.14 inches for a current of ten amperes. Hence, the core should consist of two concentric annular rings. The inner ring, with an inner diameter of 0.08 inches and an outer diameter of 0.28 inches, is made of molypermalloy, and the outer ring, with an inner diameter of 0.28 inches and an outer diameter of 0.6 inches, is made of ferrite. The resultant inductance may be computed as follows:

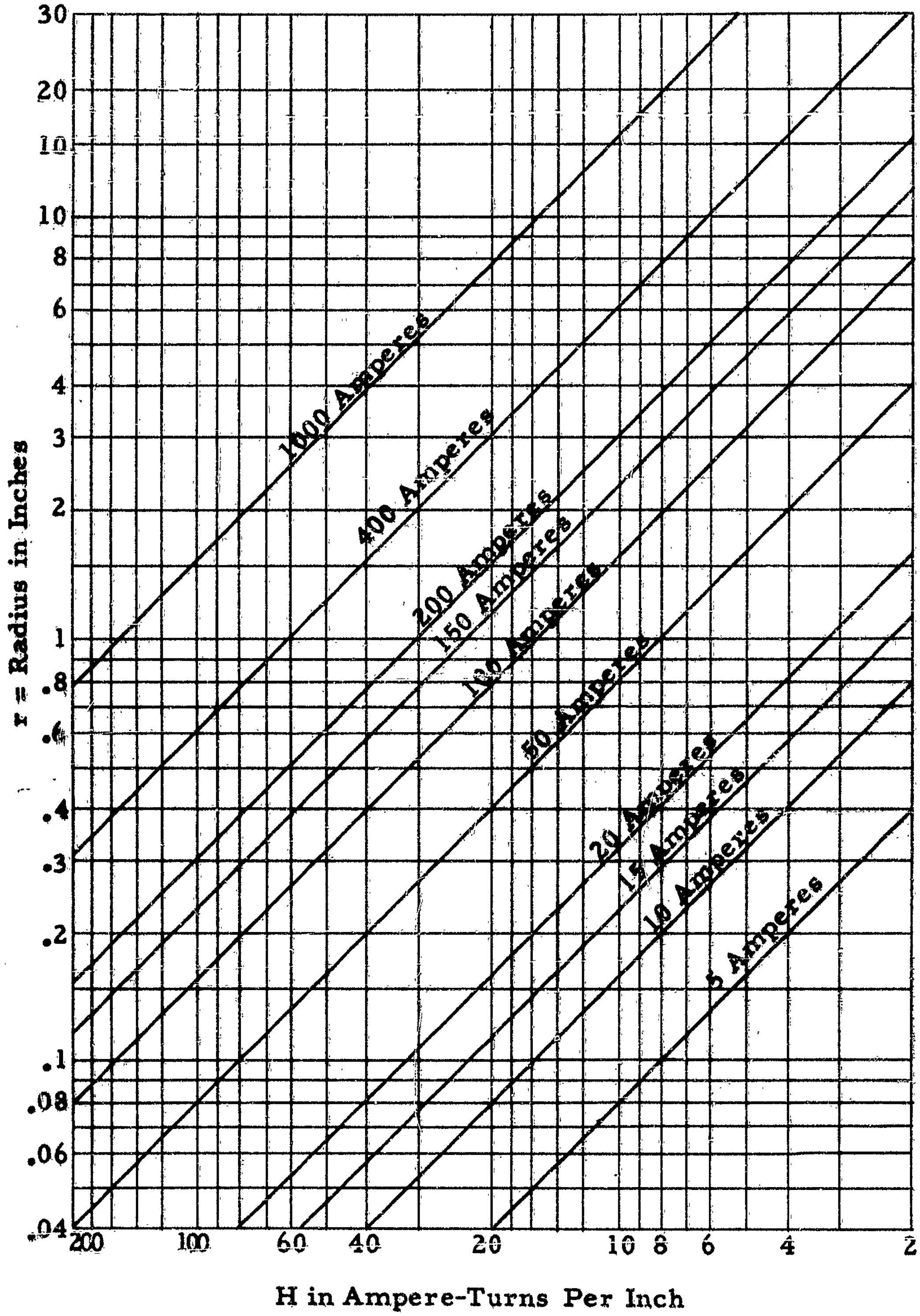


Fig. 3.1.1.6-J Relation Between Radius From Center of Cylinder and Magnetic Field Intensity for Different Currents

$$L' = 0.00508 \left(\mu_{d_E} \ln \frac{0.3}{0.14} + \mu_{d_C} \ln \frac{0.14}{0.04} \right) \qquad (3-18)$$

The quantity μ_{d_E} varies from 135 to 105 in the region considered. An average of 120 may be used. The quantity μ_{d_C} varies from 105 to 97. An average of 101 may be used. With these values the inductance, L', is 1.11 microhenries per inch.

In choosing a magnetic core for an inductor, three factors must be considered: permeability, saturation, and losses. When inductors are used in networks that must pass only direct or very low frequency currents, the losses need not usually be considered. The losses always increase with frequency, and when all high frequencies are to be suppressed, losses are not objectionable. They are just additional aids in attenuating the interfering currents. But in harmonic-suppression filters or other networks that must pass radio frequencies, low-loss materials must be chosen. Figure 3.1.1.6-G shows that the ferrites* have the highest initial permeability and the lowest losses at 200 kc. However, they reach saturation very quickly, and their losses increase very rapidly at frequencies above 5 mc. Special ferrites have been produced that are useful at frequencies as high as 100 mc, but their permeability is much lower. Cores of powdered iron or special alloys such as molypermalloy have very low losses even at very high frequencies. They are not saturated easily, but their initial permeabilities are not very high. They are still the best compromise available when fairly high inductances and low losses at high frequencies are required.

It is often desirable to embed coreless inductance coils in plastic material in order to protect them from shock. When doing this, the heat-conducting properties of the material become important. Powdered quartz has been found particularly useful since its large heat conductivity allows an increase of the current rating, thus permitting larger values of inductance without increase in volume.

3.1.2 SHIELDING

The purpose of a shield is to keep all radio-interference energy "bottled up" within a specified region, or to prevent all radio-interference energy from entering a specified region. The first type is used for ignition systems, motors, and other sources of radio interference. The second type is used for receivers or leads leading to receivers. Because power for control energy must always be supplied or removed from the region within the shield, and because the techniques of construction as well as the necessity for accessibility and serviceability demand that shields be made of more than one part, openings, seams, joints, or other discontinuities must always be present. The problem of constructing an effective shield has therefore two separate phases: One is the prevention of the penetration of electro-magnetic energy through the shielding wall itself, and the other is the prevention

*Ferrites are mixtures of crystalline iron oxides of ceramic-like structure. To avoid confusion with the word "ferrite" as used for metallic iron containing a fraction of one percent of carbon, the name "ferrospinels" has recently been suggested for these materials.

of leakage through the discontinuities in the shield. The second of these two problems - the proper design of the necessary discontinuities so that effectiveness of the entire shield is not impaired - requires the greater consideration and attention. Just as a chain is no stronger than its weakest link, a shield is no more effective than its poorest joint. The major portion of the following paragraphs is devoted to the principles and techniques employed in the design of discontinuities in shields for minimum leakage.

3.1.2.1 SHIELDING MATERIALS

The problem of preventing penetration through the shielding wall itself is comparatively simple. As was pointed out in Paragraph 1.8.2.2, under certain simplified conditions the ratio of the electromagnetic energy that has penetrated a shield to that entering it, expressed in decibels, varies inversely as the thickness and the square root of the magnetic permeability and directly as the square root of the resistivity. While shielding effectiveness depends on other factors, such as the impedance of the wave and the geometrical shape of the shield, in a very complicated way, this dependence can be entirely neglected provided that the three factors mentioned above are chosen large (or small) enough. This leads to walls that may be much thicker than necessary for the desired degree of shielding, but usually not thicker than necessary for mechanical reasons whenever the shield must support itself mechanically. In the absence of more detailed information about the effect of the other factors, it is suggested that the minimum thickness of shielding material be based on an absorption loss of about 33 db at 1 mc. In Appendix XVI the following expression is developed for the absorption loss in decibels.

$$\text{Absorption Loss} = 3.34\, S \sqrt{f_m \mu_r \sigma_r} \qquad (3\text{-}19)$$

where S is the thickness in mils, f_m is the frequency in megacycles per second, μ_r is the relative magnetic permeability (μ_r = 1 for all non-magnetic materials) and σ_r is the conductivity relative to copper (see Figure 3.1.2.1). Hence, the thickness for about 33 db at 1 mc is given by:

$$S = \frac{10}{\sqrt{\mu_r \sigma_r}} \qquad (3\text{-}20)$$

This leads to a minimum thickness of 10 mils for copper - a choice that has proved satisfactory in practice.

When magnetic materials are used, care must be exercised in the evaluation of μ_r because it varies with saturation and frequency. If the frequency is high enough, the permeability, even of the best magnetic materials, decreases to unity. Therefore, the shielding effectiveness of magnetic materials, such as steel, should be carefully tested before using Equation (3-19) or (3-20) with a value of μ_r obtained from low-frequency measurements. Figure 3.1.2.1 shows the minimum thickness of shielding recommended for several common shielding materials together with their relative conductivities. The value for steel is conservatively based on a permeability of unity. In most cases this thickness is insufficient from a mechanical point of view, but it is the suggested minimum when there are no mechanical considerations.

Metal	Relative Conductivity, σ_r	Minimum Thickness in Mils
Aluminum	0.61	13
Brass	0.25	20
Copper	1.00	10
Magnesium	0.37	16.5
Silver	1.05	10
Steel	0.035 -0.16	25 - 55
Tin	0.15	26
Zinc	0.29	18.5

Fig. 3.1.2.1 Minimum Thickness Recommended for Shielding for Metals Having Conductivities Shown

The shielding effectiveness of a solid metallic wall increases with frequency (except possibly for magnetic materials). Therefore, measurements of this effectiveness need to be made only at the lower frequencies. In fact, it has been found in practice that, if a particular material and thickness are satisfactory below 20 mc, they will be satisfactory above that frequency. This condition may be vitiated, however, by the effect of seams, joints, or other discontinuities. For such discontinuities the opposite is true: Their shielding effectiveness decreases with frequency so that joints which are entirely satisfactory at low and medium frequencies may be quite "leaky" at high, very high, or ultra-high frequencies.

Instead of solid metal walls, meshes of metallic wires are sometimes used for shielding purposes. The attenuation of an electromagnetic wave in a mesh is considerably less than that in a solid screen. Therefore, the principal shielding action of a mesh is due to reflection. Tests have shown that mesh with 50 percent open area and sixty or more strands per wavelength introduces a reflection loss very nearly equal to that of a solid sheet of the same material. For this to be true, it is necessary that the mesh be so constructed that the individual strands are permanently joined at their points of intersection by some kind of fusing process so that good permanent electrical contact is made. Figure XVI-3 in Appendix XVI shows that reflection loss of a solid shield decreases with frequency. In addition, the reflection loss in a mesh depends on the number of strands per wavelength, and since the wavelength decreases with frequency, the shielding effectiveness of meshes decreases with frequency faster than indicated by Figure XVI-3. Since the reflection loss is also that portion which is most affected by the impedance of the wave, and hence the configuration of the source, it is best to make careful tests whenever meshes are to be used for shielding rather than to rely on theoretical considerations. More details on the reflection loss, absorption loss, the effect of the wave impedance, and the use of mesh shields are found in Appendices VIII and XVI.

3.1.2.2 OPENINGS IN SHIELDS

A case designed to completely enclose a unit such as a receiver or motor must have openings and other discontinuities for the following purposes: to pass power, control, and output leads; to allow access for maintenance and servicing; to aid the ease of manufacture and assembly; and to permit proper ventilation, drainage, and

heat transfer. Of these only the considerations of ventilation, i.e., the circulation of air, and drainage of any condensed moisture require actual openings. All the other requirements can be satisfied with temporary, semi-permanent, or permanent seals.

Leakage of electromagnetic energy through actual openings may be minimized by controlling either the size or the shape of the holes. The amount of electromagnetic energy that can escape through a hole in a shield is roughly proportional to the size of the hole provided that its dimensions are small compared to a wavelength. This means that the leakage can be kept negligible simply by making the holes sufficiently small. When the only purpose of the hole is the drainage of condensed moisture, a small number of very small holes of no more than 1/8-inch diameter is usually sufficient, and the leakage through these holes is negligible except in the case of extremely powerful interference sources such as ignition systems or radar modulators. For proper ventilation, larger openings are required. Such openings must then be covered with fine-mesh copper screen which must be soldered or welded along a continuous line around the edge of the opening. The type of mesh must be chosen in accordance with the principles explained in the previous paragraph. Since a mesh for effective shielding action has rarely more than 50 percent open area, and frequently less than that, the size of the opening must be correspondingly increased for effective ventilation. If the mesh must be easily removable, it must be attached with screws or bolts in sufficient number to insure a high pressure contact along a continuous line completely around the edge. Here, as in other joints, maintenance constitutes the largest problem. The contact surfaces must be thoroughly cleaned each time the mesh is screwed back into place.

An alternative to reducing the size of the openings, either by making the holes themselves sufficiently small or by covering them with metallic meshes effectively making many small holes out of one big one, is to design the shape of the openings in such a way that the escape of electromagnetic energy through them is prevented. The most effective way of doing this is to surround the openings by protruding sleeves which, effectively, convert the openings into wave guides. To be considered a wave guide, the length of the sleeve should be at least three times its longest cross-sectional dimension, or three times its diameter if of circular cross-section.

Wave guides act like high-pass filters. They pass without attenuation (neglecting losses in the walls and dielectric) all frequencies above the cut-off frequency and attenuate all frequencies below the cut-off frequency. Since a particular wave guide permits many modes of transmission of electromagnetic waves through it, there are many cut-off frequencies associated with each guide, one for each mode of transmission. The lowest cut-off frequencies for rectangular and circular wave guides are given by the following expressions:

$$f_c = \frac{5900}{b} \tag{3-21a}$$

$$f_c = \frac{6920}{d} \tag{3-21b}$$

where f_c is the lowest cut-off frequency in megacycles per second, b is the longer inside dimension of the rectangular guide in inches, and d is the inside diameter of

the circular guide in inches.

In a wave guide, the attenuation per unit length for a wave below the cut-off frequency of the wave guide, neglecting dissipation, is given by the following expression:

$$\alpha = 0.00463 \; f \sqrt{(f_c/f)^2 - 1} \tag{3-22}$$

where the attenuation, α, is in decibels per inch, and f is the frequency in megacycles per second. When the frequency is less than one tenth of the cut-off frequency, a good approximation, independent of frequency, may be obtained in the following form:

$$\alpha = \frac{27.3}{b} \tag{3-23a}$$

$$\alpha = \frac{32}{d} \tag{3-23b}$$

for rectangular and circular guides, respectively. Figure 3.1.2.2 shows a plot of

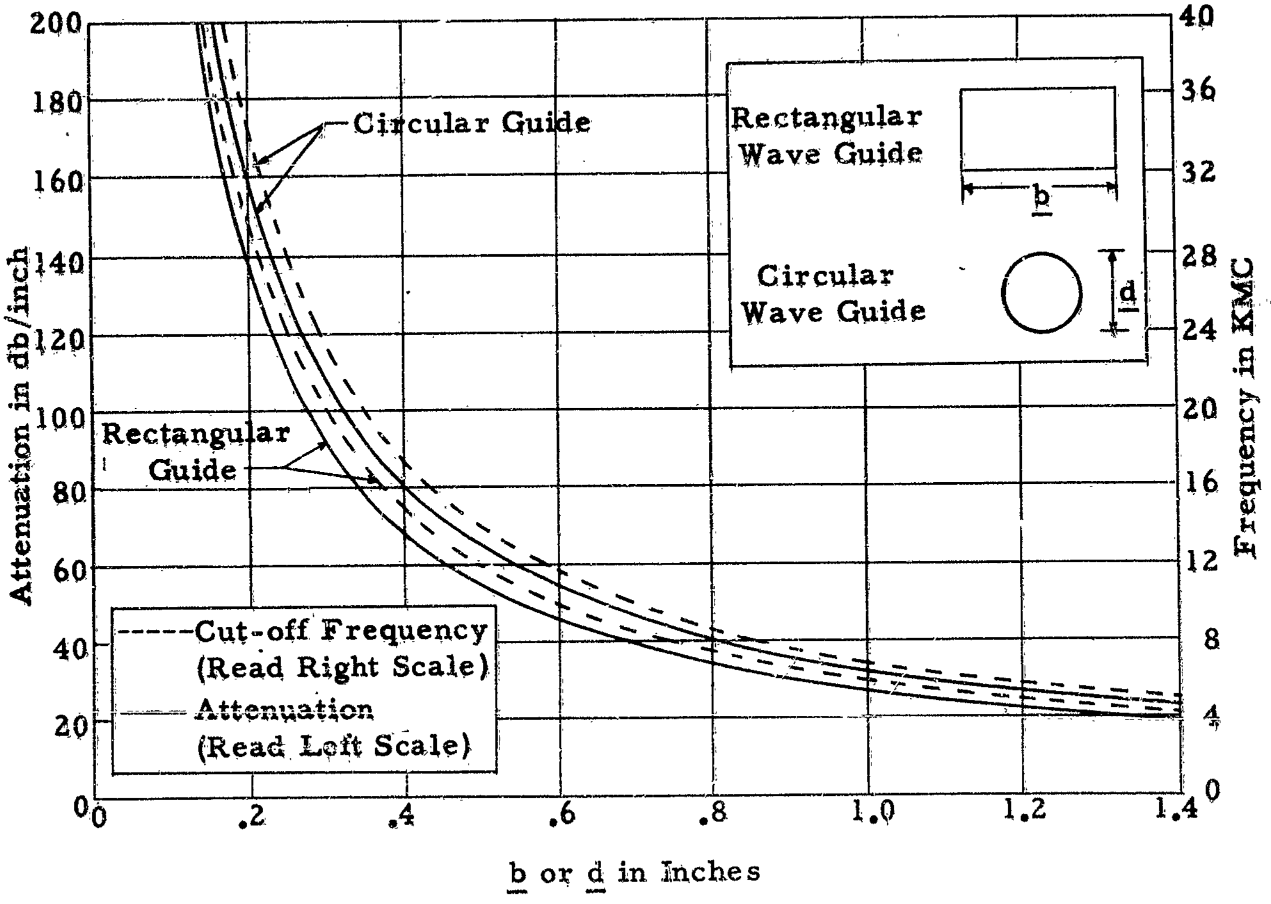


Fig. 3.1.2.2 Cut-off Frequency and Attenuation of Wave Guides

Equations (3-23a) and (3-23b). From these curves the attenuation per unit of length (decibels/inch) may be read directly for any size of rectangular or circular wave guide. The same figure also shows the cut-off frequency (Equations (3-21a) and (3-21b)) in each case, so that the limit of validity can be determined from the same figure.

> For example, if 70 db are required up to at least 600 mc in a circular wave guide, the inside diameter must be no more than 1.15 inches based on a cut-off frequency of 6000 mc. For that size, the attenuation is 27.2 db per inch, so that about 2.5 inches of length would apparently give 70 db of attenuation. But remembering that the minimum ratio of length to diameter is three, a length of three inches should be chosen.

It is seen that, since the minimum length is 3b or 3d, the minimum attenuation is always at least 82 or 96 db for all frequencies up to about one-tenth of the cut-off frequency. A picture of an end-cap for a direct current motor whose openings for ventilation were designed in accordance with these considerations, which proved very effective in actual tests, is shown in Figure 3.2.1.1.8-A.

3.1.2.3 JOINTS

When it becomes necessary to join together several parts of a complete shield, the first consideration must be to keep the number of such joints down to the absolute minimum. In practically all cases where joints are necessary to allow ease of manufacture and access for maintenance, the shield should be constructed of no more than two parts having only one joint. In some cases it is most desirable to make a permanent joint by welding or soldering and sacrifice accessibility. This is done, for example, with ignition coils which are sealed permanently in a metallic shield. When the coil fails, the shield and coil are considered expendable and are replaced as one unit. This extreme solution of the problem of joint design cannot, of course, apply to more complicated pieces of equipment that may require frequent servicing.

When joints are made, the most important requirement is that a continuous metal-to-metal contact be maintained along a continuous line. When the pressure is maintained by means of screws or bolts, a sufficient number must be used to assure high unit pressure even at the points most distant, i.e., farthest away, from any screw or bolt. Lack of stiffness of the mating members is an important factor in producing distortion of the mating surfaces, a condition which results in bulging and insufficient pressure for good electrical contact. Flange and cover-plate joints should be made circular wherever possible because of the ease with which the surface can be machined either plane, grooved, or tapered. An example of a good shield design with a circular joint is shown in Figures 3.1.2.3-A and B which show a photograph of a generator-regulator shield assembly open, and a similar assembly, closed. A retaining band is used to maintain high unit pressure all around the joint.

A modification of the taper or wedge cover-plate design is shown in Figure 3.1.2.3-C. Here the shape of the mating members assures positive contact along two continuous lines. For thinner shielding materials, when screws cannot be used because the members are not stiff enough to maintain high pressure between screws, the "paint can" cover illustrated by Figure 3.1.2.3-D gives good results. It can be used only for shields that need not be designed for strength and rigidity.

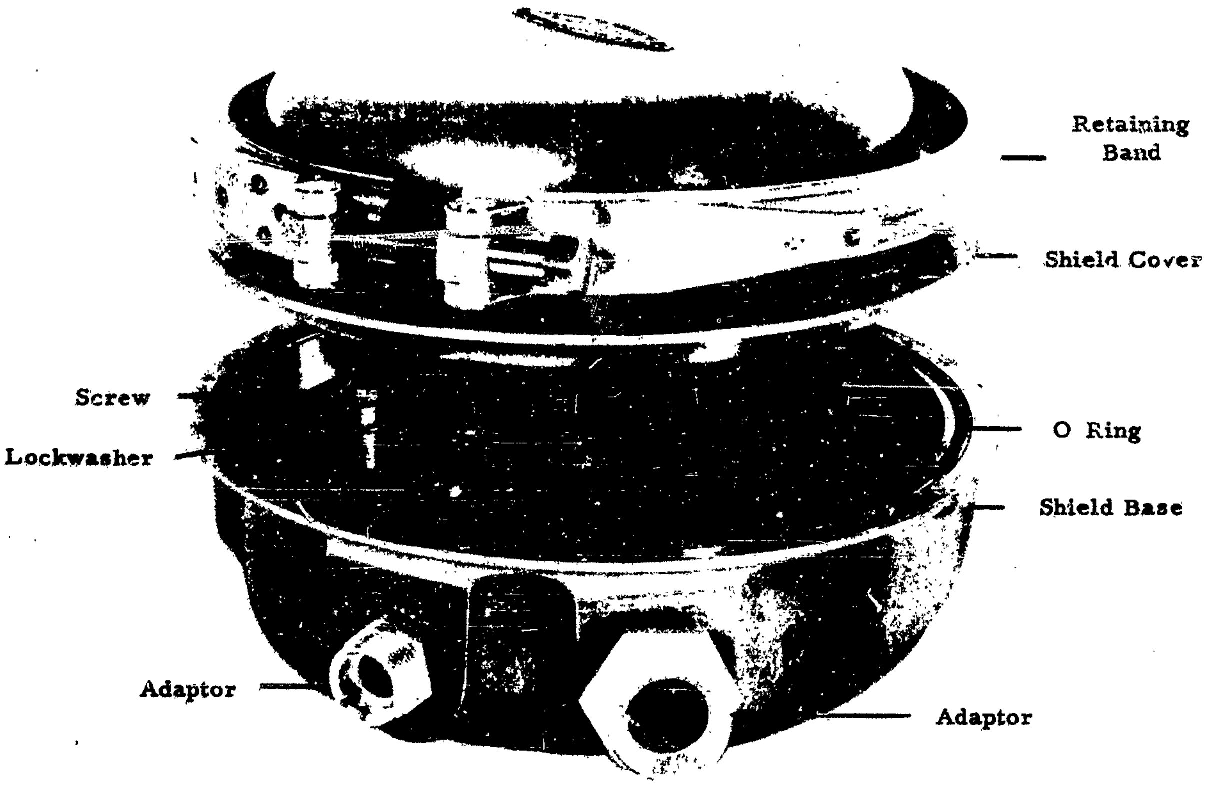


Fig. 3.1.2.3 - A Generator Regulator Shield Assembly Open

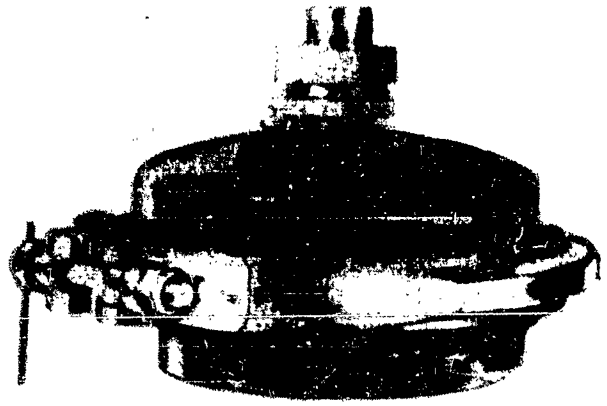


Fig. 3.1.2.3 - B Tachometer Shield Assembly Closed

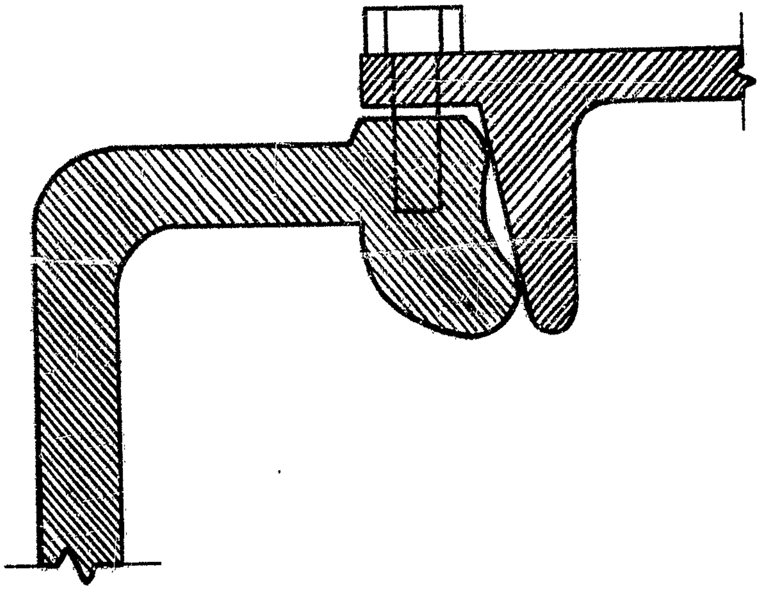


Fig. 3.1.2.3-C Design of a Tapered Cover Plate.

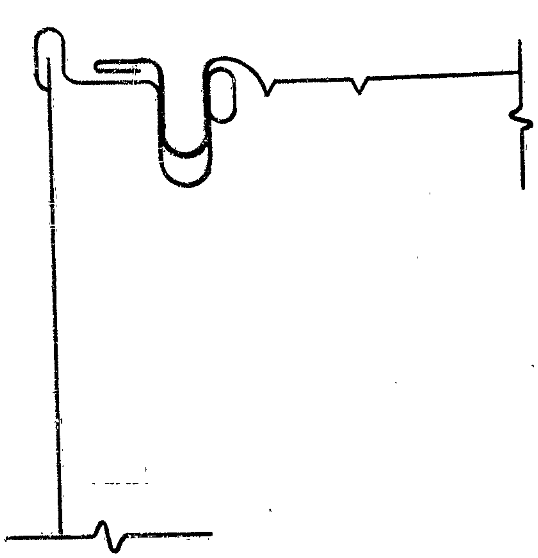


Fig. 3.1.2.3-D "Paint Can" Cover Plate

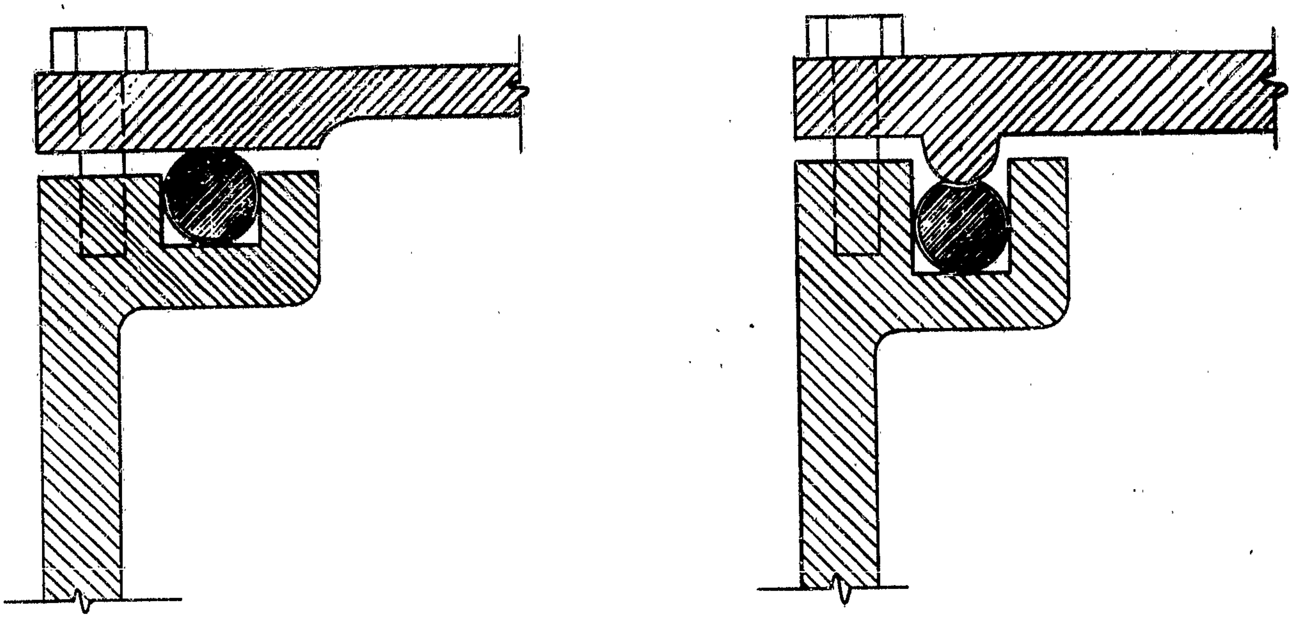


Fig. 3.1.2.4-A Cover Plates with Conductive Gaskets

3.1.2.4 CONDUCTIVE GASKETS

Failure of the flange and cover-plate designs to provide continuous line contact has necessitated the use of various types of conductive gaskets at the interfaces of mating surfaces. Two examples of designs of cover plates using conductive gaskets are shown in Figure 3.1.2.4-A.

The use of conductive gaskets in any shielding assembly is an admission of the inadequacy of the joint design because if continuous line contact at all mating surfaces were attained, there would be no need for gaskets even for providing moisture and gas seals at the joint interfaces. Yet, conducting gaskets are used in many types of shielding of present design and consequently must be discussed in some detail.

Any gasket must have a degree of compressibility if it is to conform to the mating surfaces. The degree of conformance is largely dependent on the available pressure and the compressibility of the material of which the gasket is made. It can safely be said that there is no one ideal gasket material for all purposes, but the successful employment of any type of gasket can be greatly improved if its use is properly taken into design consideration at the outset. The introduction of a gasket at a later date may not be as highly successful.

Seals for radio interference reduction generally fall into one of two classes, the multiple-contact type or the continuous-contact type. A typical example of the former is a Neoprene-impregnated screen, and of the latter, the foil-wrapped gasket. Since gaskets are essentially interposed in the shielding system, they must satisfy the requisite degree of attenuation. For most metals, this can be calculated from depth of penetration formulas. In gaskets this is difficult except for the foil-wrapped type, where it is obvious that the foil thickness must be that required for effective penetration loss. In the multiple-contact type, the greater the number of contacts, the more effective the gasket becomes.

Conductive gaskets ordinarily are used only on the flange and cover-plate type of joints. The various types of gaskets either now in use or proposed for use are:

(a) Metal screen impregnated with Neoprene (excess Neoprene removed from wire-mesh surface by use of abrasives).

(b) Metal foil over Coroprene core.

(c) Wire sleeving over Neoprene core (for use in slotted flange and cover-plate design).

(d) Sprayed-metal Neoprene gaskets.

(e) Aluminum-Magnesium alloy crystals suspended in Dow Corning No. 4 Compound.

(f) Serrated washer-type metal gasket.

(g) Metal-brush type (for use in slotted flange and cover-plate design).

(h) Aluminum tubing filled with a Neoprene core (for use in slotted flange and cover-plate design).

(i) Compressed knitted wire mesh of copper, monel, or other metals.

A brief discussion of the above types of conductive gaskets, based on the experiences of the Military Services and on performance tests made by various equipment manufacturers follows. The letters refer to the gasket types described above.

The wire sleeving over a Neoprene core (a), the aluminum tubing filled with a Neoprene core (h), and the knitted wire mesh (i), for use both in flange and cover plates of slotted design, are the most promising of the several suggested types for providing interference-free service.

All gaskets that depend on multipoint contact for shielding, namely, the wire mesh, Neoprene-impregnated (a), the serrated metal washer (f), the metal-brush type (g), the Al-Mg alloy crystals (e), and the knitted wire mesh (i), are effective in proportion to the density of the points of contact, or in other words, to the degree to which line contact is approached.

The use of finely divided Al-Mg alloy crystals, while effective in providing multipoint contact, which closely approaches line contact provided the space between mating surfaces is not greater than the crystal size, is not recommended by the Military Services because of the danger that might be encountered should the metal particles inadvertently come in contact with the surfaces of bearings.

The metal-foil covered gaskets (b), while suitable for use on flat surfaces, have caused trouble on curved surfaces due to buckling or breaking of the foil.

Sprayed-metal Coroprene gaskets (d) were found by test to be inferior to other gaskets because of the porosity of the coating.

The knitted wire mesh (i) is far superior to any gasket made of woven mesh. It can be made from any metal or alloy that can be drawn into wire. It can also be combined with a sealing medium such as Neoprene when hermetic sealing, as well as shielding, is required.

Conductive gaskets can be made in practically any size or shape to fit the equipment with which it is to be used. The resiliency and density of the gasket material can also be varied according to the requirements of each application. A well-designed gasket insures good all-around contact even with appreciable unevenness of the mating surfaces or warping. Special machining for a close fit is not required.

Shielding gaskets for use in the covers of sheet-metal chasses and cabinets are available in round strips, strips with one or two fins, and double strips with a connecting web to meet varied installation requirements. Typical examples of the use of these types are shown in Figure 3.1.2.4-B.

When conductive gaskets are used with flange joints, the gasket should be thicker than the depth of the groove that holds it to insure complete contact, as shown in Figure 3.1.2.4-C. It is also important that the gasket be located inside the flange

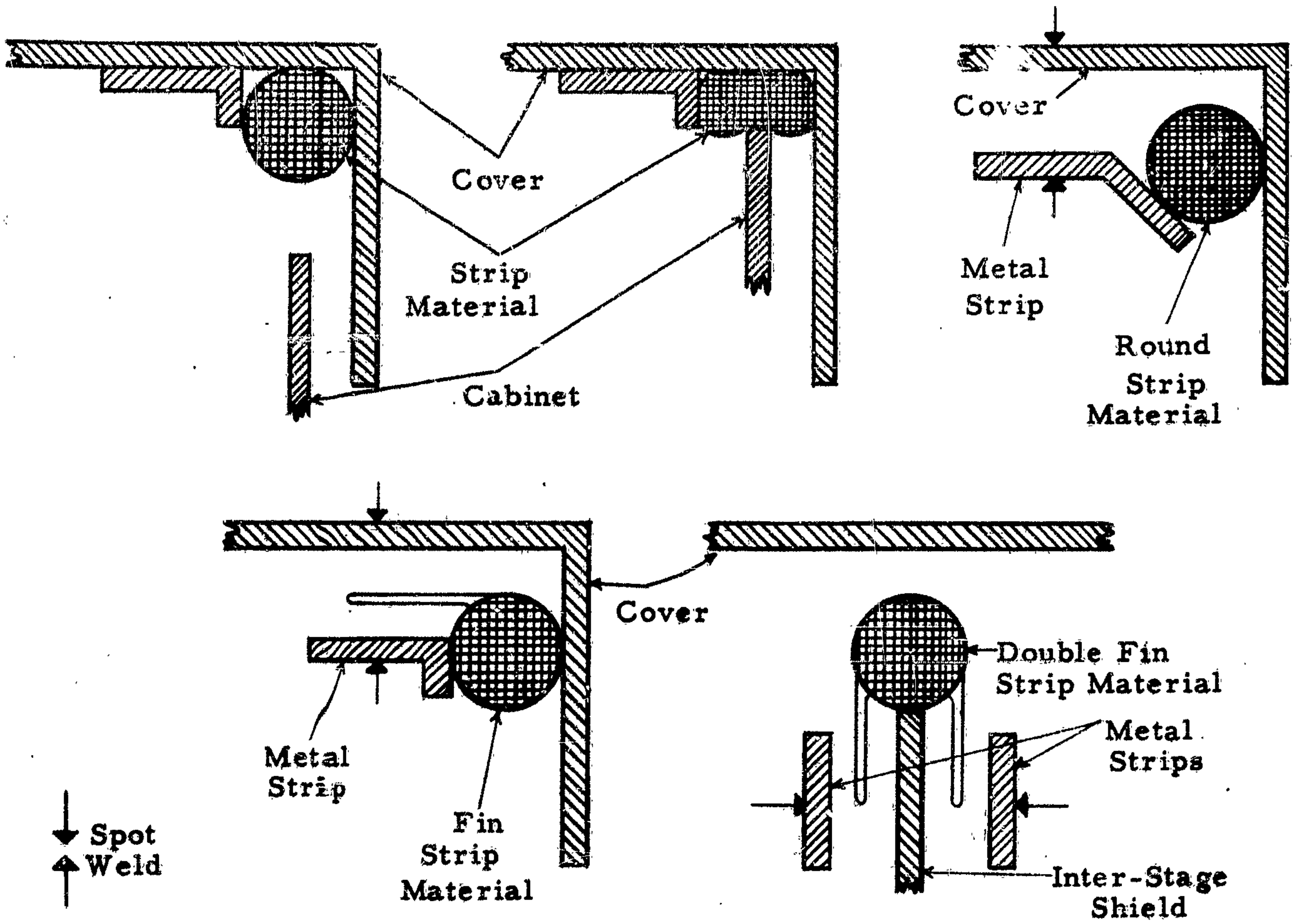


Fig. 3.1.2.4-B Use of Round, Single-Fin and Double-Fin strips of Conductive Shielding Gaskets

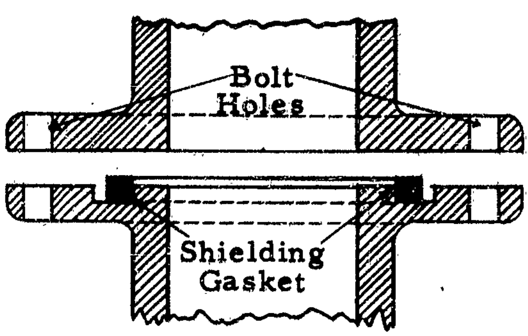


Fig. 3.1.2.4-C Use of Conductive Gasket with Flange Joint

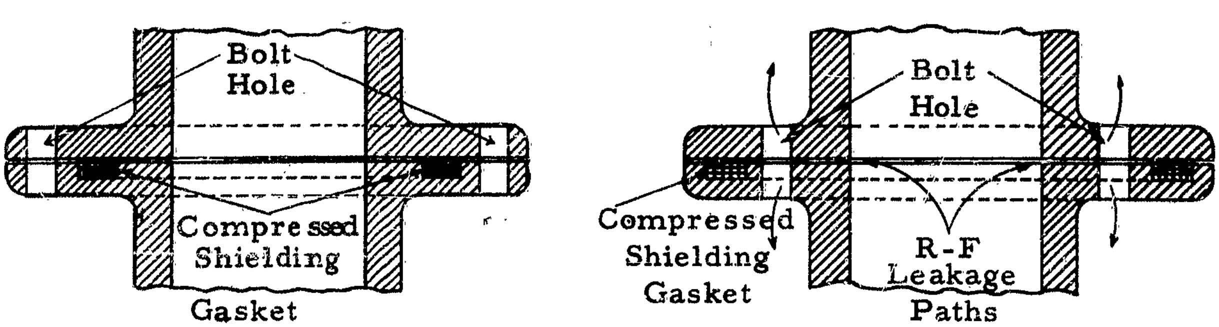


Fig. 3.1.2.4-D Proper Location of Shielding Gasket

bolts to prevent leakage through the bolt holes. This is illustrated in Figure 3.1.2.4-D.

In choosing the material for the conductive gasket, the electrical conductivity is only one of many factors to be considered. For example, gaskets of monel metal have been found to shield as effectively, if not better than, gaskets made of silver-plated copper. In addition, monel metal gaskets have greater resiliency for the same volume of metal and are much more resistant to corrosion. In fact, the ability to resist corrosive action - atmospheric or galvanic - is often as important as the electrical conductivity.

In spite of the many advantages of conductive gaskets, it must be remembered that a joint with a gasket can never produce as good a shield as a good joint with direct metal-to-metal contact. Tests have shown that a gasketed joint may have eight times the impedance of a similar metal-to-metal joint. This is due to the fact that a gasketed joint comprises two joints in series and may have a useful contact area of roughly one quarter of its apparent contact area. But good direct joints are very difficult to obtain, and still more difficult to maintain, and in general gaskets may improve the performance of average-to-poor joints.

3.1.2.5 JOINT PRESSURE

The pressure exerted on the mating surfaces of any joint in a radio interference shield must be of sufficient magnitude to maintain line contact of the mating metals and to provide a low-impedance path at their interface. The subject of pressure as a function of the quality of electrical contact is discussed in greater detail in Paragraph 3.1.2.6.

The problem of joint pressure in the flange and cover-plate type of shielding joint is of great importance because of the comparatively large and irregular surface area that must be joined together. Many joints of this type, as designed in the past, have not been satisfactory with regard to prevention of radio interference because of

(a) failure of the joints to provide continuous line contact because their mating surfaces were not machined sufficiently plane. In some instances, it was necessary to scrape and lap the mating surfaces of this type of joint in order to achieve the necessary line contact,

(b) failure of the joints to provide continuous line contact because their mating surfaces became "waved" as a result of warping when pressure was applied to the joint assembly by screws or bolts spaced at intervals along the periphery; also, similar trouble has been encountered as a result of thermal changes in the joint assembly. The causes for this trouble can be traced to a lack of stiffness of the flange and plate edges and to excessive distances between the points of application of pressure, and,

(c) failure of the joint design to provide suitable means for maintaining constant pressure at the mating surfaces, especially during operation under service conditions when vibration and heat tend to loosen the fastening mechanism and reduce the joint pressure.

3.1.2.6 CONTACT IMPEDANCE

In the design of joints in radio interference shields, there are conflicts between mechanical, electrical, and chemical requirements, all of which need careful consideration in view of their close interrelation. A good design, as always, is one that achieves the best possible compromise among the various factors involved.

The impedance between two conductors that are pressed together is in most cases largely resistive; however, it can be inductive or capacitive depending on the nature of the joint and the wave length involved. Therefore, the general term impedance is used to include the cases of the highest frequencies and dimensions. In any case, the total impedance across two conductors in contact must be carefully considered by the designer. It consists of

(a) The impedance of a major part of the first conductor, in which the current density is almost constant, owing to the comparative remoteness of the interface.

(b) The impedance of the remainder of the first conductor in the proximity of the contact area or areas where the current density is quite variable.

(c) The impedance of the film between the conductors at their contact area or areas. The film may consist of one, two, or three films in series. Each contact surface, in general, has a certain amount of "bound" film due to corrosion or surface treatment; also there is a certain amount of "floating" film, such as grease, not securely attached to either contact surface. However, bound films are not always tenacious, and floating films sometimes adhere firmly.

(d) The impedance of the second conductor just beyond the interface, where the current density is quite variable.

(e) The impedance of the second conductor for some distance beyond the interface, a region in which the current density is almost constant.

In most practical cases, contact impedance consists mainly of "interfacial impedance" (c) and the combined "constrictional impedance" (b) plus (d). Contact impedances are, therefore, sensitive to a number of factors, many of which do not prevail in ordinary solid conduction. The factors to be considered especially in designing shielding joints are (a) the conductivity, hardness, density, roughness, cleanliness, and corrosion-resistance of the materials, (b) the size and geometrical shape of both mating surfaces, and, (c) the pressure, that is, the force per unit area binding the surfaces together.

Since in most cases the reactive portion of the impedance is very small compared to the resistive component, it can be neglected and the effective contact resistance, or real part of the impedance, can be approximately determined by the equation:

$$R = KP^{-N} \qquad (3\text{-}24)$$

where R is in ohms, K is constant for a system, P is the pressure, and N is an exponent usually lying between 0.5 and 2.0, depending upon the nature of the contact surfaces. The determination of K and N is somewhat difficult because of their interdependency on the materials and geometry. The test data on R as a function of P may be plotted to great advantage on log-log graph paper. The resulting points approximate a straight line of negative slope N.

Logarithmic curves of R plotted as a function of P usually show a marked linearity - with erratic departure from a straight line taking place only at very low and extremely high pressures. Extremely rough or warped surfaces show a very high contact resistance at low pressure, but a rapid decrease in resistance with added pressure. Such surfaces at first may touch at only two or three points, but due to high unit pressure, these points are eventually depressed, permitting other points to make contact. Extremely smooth or lapped surfaces also show very high contact resistance at low pressure, but their decrease in resistance with added pressure is less marked. Such surfaces may, at first, engage films of corrosion, oil, air, and water, all of which are difficult to eliminate with low unit pressure.

Theory and experience indicate that the best contact surface is a slightly rough, though clean, surface. It has been noted that highly polished platinum contact points, often used in breaker assemblies, operate best after a brief run-in period. The preferred range of roughness must be determined in various cases. It should not be so rough as to cause severe air leakage around an ungasketed joint. A surface may not actually appear rough except under a magnifying glass. Instruments are available for the measurement of roughness, but it might be unwise and unnecessary to attempt to use them in production testing because once a satisfactory range of roughness has been determined, it is not difficult to specify a control standard.

In the design of shielding joints, an important question arises as to whether or not any particular flange width or area results in a minimum direct current resistance when the clamping force is constant. This question may be approached analytically. The measured resistance R of a metallic conductor is directly proportional to the length L of the conductor and inversely proportional to its cross-sectional area A. Introducing a constant of proportionality, ρ, called the resistivity of the metal, permits expressing the measured resistance as:

$$R = \rho L/A \qquad (3\text{-}25)$$

For the case in question, ρ refers to the joint itself rather than to the metal at either side. The length L refers to the thickness of the interstice between the mating surfaces, and A refers to the actual cross-sectional area, even though all of it does not effectively conduct current. Obviously, the values of ρ and L depend primarily on the nature of the mating surfaces and are independent of the area A. All three factors on the right side of Equation (3-25) are actually functions of the pressure but A, as defined for this case, changes a negligible amount with pressure. Therefore Equation (3-25) can be written as:

$$R = f(P)/A \qquad (3\text{-}26)$$

where the function, f(P), accounts for the change in ρ and L with pressure. Equation (3-24), derived empirically, accounts for the change of these two factors with

pressure by the value of the exponent N.

The problem is to determine the effect of a change in the actual cross-sectional area A on the measured resistance R when the force F holding two mating surfaces together is kept constant, all surface conditions remaining the same. Analogy to Equation (3-26) shows that K of Equation (3-24) is, in general, a function of the area A. Since the resistance is inversely proportional to A, as seen from Equation (3-26), the function K must be equal to K'/A, where K' is a true constant independent of both the area and the pressure. If F is the constant force, the pressure P is F/A. Substitution of these expressions into Equation (3-24) gives:

$$R = \frac{K'}{A}\left(\frac{F}{A}\right)^{-N} = K'F^{-N} A^{N-1} \qquad (3\text{-}27)$$

It follows then, for a constant F, that R is independent of A when N = 1, increases as A increases when N > 1, and increases as A decreases when N < 1.

This means that the flange width has no effect on the direct current resistance when the total clamping force is constant, providing the log-log curve (abscissa and ordinate having same scale) of R plotted as a function of P has a 45° slope. If the slope exceeds 45°, as in the case of very sharp or rough surfaces, such as the perforated steel-core gasket, the flange width should be small. If the slope is less than 45° (N less than 1), as in the case of all other gaskets and all surfaces not covered with metal paint, the flange width should be large. As N approaches zero, an increase in clamping force is less effective than a corresponding increase in area in reducing resistance. When N varies from 0.5 to 0, the effect of area changes from an inverse square root to an inverse proportion (as in wire conductors).

As a general design consideration for interference-free operation, it is well to keep in mind that the direct current resistance of practically all contacts is reduced by the use of wider contact surfaces; however, radio frequency impedance is not proportionately reduced, and gaskets, when used, may be less satisfactory. A narrow contact is generally a good contact because it has high unit pressure. It thus promotes good shielding if it is wide enough to pass substantially all of the radio frequency currents.

3.1.2.7 CORROSION

Corrosion is a very important factor in the choice of design of shielding joints for permanent interference-free operation. Corrosion is a chemical action or effect whereby metals are gradually disintegrated and converted to high-resistance compounds. Exposure to ordinary air results in surface corrosion on all but a few metals. Corrosion may be avoided by the extensive use of protective coatings, such as paint, varnish, lacquer, or grease, and chemical treatment.

One of the best protections for magnesium is a chemical treatment, known as Dow No. 7 or sodium dichromate treatment. This treatment results in very high resistance at electrical joints, and thus is objectionable in radio shielding applications. Moderate protection of magnesium, together with reasonable contact resistance, is afforded by another chemical treatment known as the Dow No. 1 chrome-pickle

treatment. This is a brief-dip process, long used as standard practice for the protection of unfinished parts and as a base for subsequent paint coats.

Until the advent of VHF radio equipment, Dow No. 7 treatments were not particularly troublesome. However, as a result of recent radio interference troubles, these coatings were removed at the radio shielding joints by abrasion, and light coatings of white petrolatum were substituted for whatever amount of corrosion protection they might afford. Experience with magnesium joints thus treated has shown, that with reasonable care, the sanding of joints did little harm to bearings, gears, and dielectric parts and that the petrolatum offered fair protection against corrosion without insulating the joints. Unless subjected to salt spray, these almost unprotected magnesium surfaces withstand rugged field service with no more corrosion in evidence than that usually seen on solder or lead.

Investigations of improved methods of surface treatment to solve the combined problems of corrosion and contact resistance lead to the following observations:

(a) Dow No. 1 treatment is a good temporary compromise.

(b) Various metal paints, such as "Metal-X", "Alumilastic", "Alkyd Graphite Varnish", and powdered-metal lacquers have been investigated, but none have been found to possess high enough conductivity for use in radio shielding applications. On the other hand, greases containing metal powders (such as DC-52) are good conductors, but have a serious disadvantage in most practical applications where these are adjacent rotary parts and dielectric material.

(c) The plating of nonconductors, such as gasket material, and of metals, such as magnesium and stainless steel, which have poor adhesion to platings, has shown that the plated surfaces are never quite impervious to moisture, and thus, when two dissimilar metals are in contact with an electrolyte, serious corrosion may eventually occur. Corrosion would take place rapidly in the case of nickel-plated magnesium exposed to the atmosphere. After a slight amount of corrosion sets in, most plated surfaces tend to peel off, especially when severe mechanical conditions are imposed.

(d) Metal sprays are subject to some of the same limitations as electroplating, but they may be very adherent and highly conductive.

(e) Iridite No. 14, a recently developed chromate process, provides an effective corrosion resistant finish, acceptable under many Armed Services specifications, for aluminum and its alloys. It can be applied to shields, mounting brackets, wave guides, connecting plugs, etc., without appreciably interfering with electrical contact because films of Iridite No. 14 offer low resistance to direct and to low or high frequency alternating currents. In addition, Iridite No. 14 provides an extremely tight paint-bond for either baked or air-dried paints and prevents the penetration of moisture to the base surface through the pores in the paint. Because of the above properties and its ease of application, Iridite No. 14 is replacing electrolytic anodizing in many applications.

Galvanic corrosion may occur when two or more kinds of metal are united in one assembly in the presence of moisture. In such galvanic couples, the corrosion of one metal is accelerated while the other metal corrodes less or not at all.

It is often necessary to combine unlike metals because of shortages and of various design requirements. Combinations such as silver and platinum, copper and monel, cadmium and steel, are known by experience to be quite compatible. New combinations of metals may be considered, in which cases little corrosion data may be available. Fortunately, it is still possible to anticipate, and thus avoid serious galvanic corrosion, without resorting to lengthy field and laboratory tests.

The first principle to observe is that joined metals should lie close together in the electromotive force series. (See Paragraph 3.1.3.1). Thus, magnesium should not be joined to brass or nickel as these combinations cause excessive corrosion of the magnesium. Certain aluminum alloys combine harmlessly with magnesium.

If the first principle cannot be strictly followed, then the second principle to observe is that the joined metals should be of such relative sizes that the attacked metal is the more abundant. For example, iron bolts on magnesium castings are fairly satisfactory, whereas aluminum rivets on brass plates are quite unsatisfactory. In the first instance, it is wise to cadmium-plate the iron and to chrome-pickle the magnesium. A third principle to observe is that joints should be kept tight and well coated in order to bar the entrance or exit of liquids and gases. A galvanic cell is powerless without moisture. It is also enfeebled (polarized) when the electrodes are coated with gas that cannot escape.

A fourth principle, often very useful, is generally inapplicable in the case of radio shielding joints. This principle requires the insulation of the joints wherever possible. Although solid insulation may be out of the question, there is no real objection to the use of semisolids such as petrolatum, which can insulate only those portions of the joints that cannot make contact anyway. Moreover, any protruding metal particles that meet are incapable of producing galvanic currents so long as moisture is kept from them by the semisolid. Naturally, in preventing corrosion, the contact impedance is kept low for a long period of time.

While the problems of corrosion and contact impedance are not yet fully solved, and until further research provides other materials with which to raise corrosion resistance while lowering contact resistance, the designer may well consider the use of bare metal surfaces for shielding joints or chrome-pickled surfaces upon which is applied an inhibiting grease, having low surface tension and suitable chemical properties as discussed above.

3.1.2.8 SOLDERED JOINTS

The corrosion problem is ever present in soldering. Corroded surfaces are difficult to solder, and soldering usually results in corroded surfaces, unless resin is used as a soldering flux. Three types of fluxes, namely, the chloride, organic acid (waxes), and organic base types, are all corrosive, differing in their rate of attack rather than in the end effects. The grease-paste emulsion of salt or acid content is little better than its fluxing ingredient, although it may be more conveniently

applied. The cooling grease may limit the travel of the liquid flux, but it does not prevent internal corrosion of the affected parts, an action which may proceed even in the absence of air or external moisture. Resin, on the other hand, is noncorrosive because it is a solid that is quite impervious to liquids and gases. For this reason, electrical joints between small clean metal parts are usually made with resin flux which may be applied as a core within solder wire.

Even a good soldered joint is likely to exhibit an appreciable contact resistance. Grade A solder has a conductivity of about 12.2 percent. Thus an ideal soldered joint is never as good a conductor as a brazed or welded joint.

3.1.2.9 MAINTENANCE

The matter of cleanliness is perhaps the most important, yet least understood, factor in a good joint. An article may appear physically clean when wiped with a "not-too-dirty"cloth. It may be chemically clean when subjected to alkaline cleansers, solvents, or chromic acid. However, metal surfaces are not electrically clean unless they make good contact, a condition exceedingly difficult to prescribe or detect.

The wiping blades of a switch may be electrically clean although coated with grease. They may be electrically dirty when bearing an invisible film of oil. Magnesium may be electrically clean when chrome-pickled (Dow No. 1 treatment) and electrically dirty when polished with a fine abrasive cloth a few hours previously. Thus, the use of certain greases between mating surfaces may provide excellent enduring contacts because they inhibit corrosion, exclude foreign matter, and have low surface tension, permitting the intimate association of adjacent metal surface particles.

The great difficulties in maintaining a joint electrically clean necessitate a continued search for types of joints that will maintain their electrical properties unchanged with time and repeated use. Improper or insufficient maintenance of shielding joints is one of the largest single causes of radio interference, observed in the field, in equipment that was found satisfactory in the laboratory. Many times, the cleaning of a mating surface, or an additional turn of a screw that seemed tight but was not, is all that is required to eliminate a major source of interference.

3.1.2.10 RIGID CONDUIT

Rigid conduit is rarely used for shielding purposes only, since damage to it by enemy gunfire requires too time-consuming repairs or entire replacement. When the use of rigid conduit is necessitated for other reasons, its incidental shielding properties may, of course, be utilized. Rigid conduit is usually made of aluminum or aluminum alloy having a wall thickness of at least 22 mils. A solid aluminum wall introduces an attenuation due to absorption of 2.6 db per mil to a plane electromagnetic wave at a frequency of 1 megacycle. Thus, 22 mils of aluminum gives an attenuation of about 57 db and this value may also be used, approximately, for a cylindrical shield. To this must be added the losses due to reflection, which depend on many factors, such as the number and position of the wires within and the shape of the conduit. The attenuation due to absorption increases as the square root of the frequency, so that at 100 megacycles the attenuation is at least 570 db. Thus, it is

seen that any rigid conduit that satisfies all mechanical requirements usually is also adequate for shielding, except possibly for shielding very strong interfering currents of low frequencies. In this last case, it may be necessary to require conduit walls somewhat thicker than those demanded by purely mechanical considerations.

Bends in rigid conduit do not affect its shielding properties provided that the metal is not damaged in any way during the process of bending. Fittings at the end of the conduit require special considerations if shielding effectiveness is not to be impaired. These will be discussed later in Paragraph 3.1.2.12.

3.1.2.11 FLEXIBLE CONDUIT

Since flexible conduit is relatively heavy and expensive, its use should be held to a minimum. Yet, there are conditions that require the use of flexible conduit, and these conditions exist quite frequently in connection with radio interference shielding. Flexible shielding conduit must be used whenever:

(a) the ends of the conduit have an appreciable relative movement, such as the conduit leading to shock-mounted equipment,

(b) shielding conduit is required at removable plugs,

(c) conduit is required at equipment subject to frequency removal, and

(d) conduit is required in the engine-section wiring subject to large vibrations.

The most important examples of the use of flexible conduit for radio interference shielding are the antenna lead-in of receivers, the wiring between different units of radar systems, and the leads to the spark plugs in ignition systems.

Flexible shielding conduit is made of strip metal formed either into spiral bellows or into some other kind of spiral that allows interlocking of adjacent strips. It may be either soldered at the seams or allowed to provide sliding action between turns. For more effective shielding, it may be covered with one or more layers of woven metal braid. One or more layers of metal braid may also be used alone without any metallic tubing inside. If so, stiffness is provided entirely by the wires carried inside the braid. It is obvious from the method of construction that flexible shielding is relatively "porous" as compared to metal tubing. It is found, as would be expected, that for the same weight and kind of metal a seamless metallic tube has greater shielding effectiveness than a flexible one.

Leakage of electromagnetic energy from flexible conduit is of two distinct types: the penetration through the metal, called "penetration-leakage", and the escape through breaks, joints, or openings, called "opening-leakage". Penetration-leakage decreases with frequency while opening-leakage usually increases with frequency. Both are present simultaneously, but the first is negligible at higher frequencies while the second is usually negligible at lower frequencies. Total leakage plotted as a function of frequency usually follows a curve as shown in Figure 3.1.2.11-A. The slope at low frequencies is determined by the thickness of the metal, the slope at high frequencies by the size and shape of the openings, and the position of the minimum, which is characteristic of practically all flexible conduits, is determined

by the design details.

The bellow construction (usually of soft brass) with soldered seams really has no openings and could, therefore, be considered "electrically tight". However, since the conductivity of solder is much less than that of brass, the shielding effectiveness is greatly reduced. Interlocked flexible metal hose, made by winding a suitably formed metal strip on an arbor and folding the edges in, as shown in Figure 3.1.2.11-B, has been widely used commercially in the manufacture of flexible tubing for shielding conduits. This construction provides three parallel metal-to-metal sliding joints between adjacent convolutions. Properly made, this type of hose provides excellent flexibility, long life under vibration, considerable ruggedness, and a very substantial degree of shielding. Compared with soldered-convoluted and seamless corrugated types of flexible metal hose, it has definite advantages with respect to shielding properties at the lower frequencies. This follows because, with the interlocked construction, it is possible to employ a much heavier gauge metal and still retain flexibility. Furthermore, the strip is folded over in such a manner that the wall of the hose is composed of four thicknesses of metal, except between convolutions.

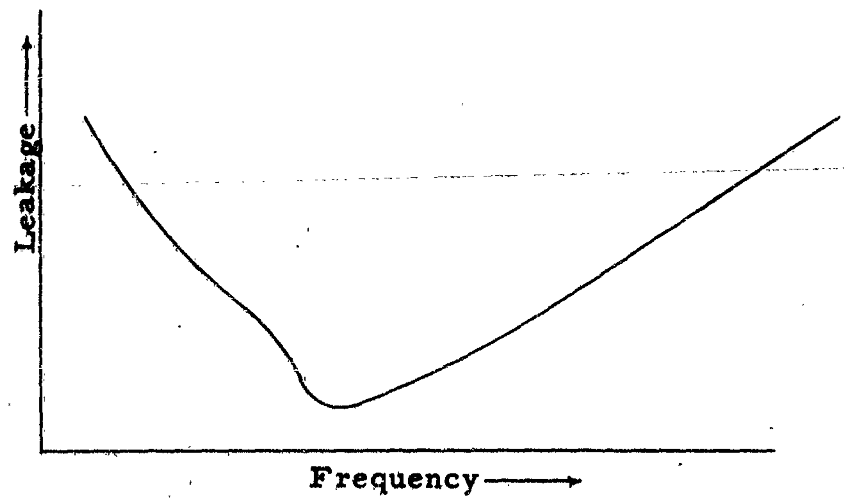


Fig. 3.1.2.11-A Leakage from Typical Flexible Conduit

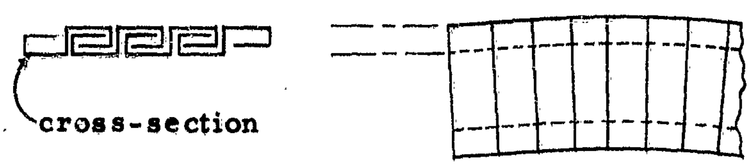


Fig. 3.1.2.11-B Construction of Interlocked Flexible Metal Hose

The interlocked construction, however, depends for its effectiveness on the attainment of good electrical contact between adjacent convolutions. If the contact is poor, relatively large opening-leakage results, which impairs the shielding effectiveness at the higher frequencies. A seamless hose or a properly made soldered-convoluted hose, on the other hand, does not suffer from this drawback. Because of its excellent mechanical characteristics and low frequency shielding superiority, interlocked hose is attractive to the designer of flexible shielding conduit. The only

real problem is that of obtaining and maintaining (throughout the useful life of the conduit) good electrical contact between convolutions.

In manufacturing interlocked hose, aluminum strip is sometimes employed. Aluminum is desirable from a weight standpoint, particularly for aircraft applications. But aluminum is a notoriously poor contact metal. Insulating films form on aluminum surfaces almost immediately upon exposure to the atmosphere. This is what makes soldering of aluminum so difficult and also why it is poorly suited for fabrication of interlocked hose for shielding purposes. Bronze and copper are found better in this respect. But even these tarnish rather quickly and the hose loses its effectiveness. Stainless steel of the so-called "magnetic" variety is found to be usable and is adopted commercially in a few cases. It has definite mechanical advantages over copper and bronze and is highly resistant to corrosion. But it is not a particularly good contact metal. Therefore, it is found difficult to manufacture satisfactory shielding hose from this metal.

Various coatings may be applied on the strip with a view to improving convolution contact. Silver is effective but tends to be costly. It is possible to apply tin to stainless steel strip and produce a very much improved hose. This development results in the type of ignition conduit known as HTCD, which has been widely used. Recent investigations have shown that further improvements may be achieved by substituting tin-coated cold-rolled steel for the tin-coated stainless steel commonly used. The improvement observed is apparently due to the much higher magnetic permeability of the cold-rolled steel. However, these observations are based on tests performed at frequencies from 50 kilocycles to 12 megacycles, and little is known about the behavior at frequencies above 12 megacycles.

The designer must be warned against the use of interlocked metal hose with an insulating cord packing which has been widely used for various applications. Because the insulating packing, wound into the convolutions, prevents electrical contact between adjacent strips, large effective openings are present and the shielding properties at high frequencies are very poor.

The use of tightly woven metal braid to cover the flexible hose is to be recommended when the hose alone does not give sufficient attenuation. At low frequencies (below 1 mc), about 45 db additional attenuation may be expected from one layer of braid, but only about 25 db more (a total of 70 db) from a second layer. At higher frequencies, each layer contributes approximately the same amount of attenuation, and a double layer may be expected to give as much as 90 db of additional attenuation.

The shielding effectiveness of various types of flexible conduit for frequencies from 50 kilocycles to 12 megacycles is plotted in Figure 3.1.2.11-C. Here shielding effectiveness is measured in terms of the surface transfer impedance (see Appendix XI). For proper interpretation, it must be remembered that a low impedance means good shielding effectiveness and a high impedance means poor shielding effectiveness. The construction of these types is described in Figure 3.1.2.11-D.

While a search for improved types of flexible shielding conduit with minimum weight continues, it may be said that, for those applications where a large degree of shielding is required (as in ignition and radar systems), a convoluted type of hose with two braids should be used with special attention given to good electrical contact

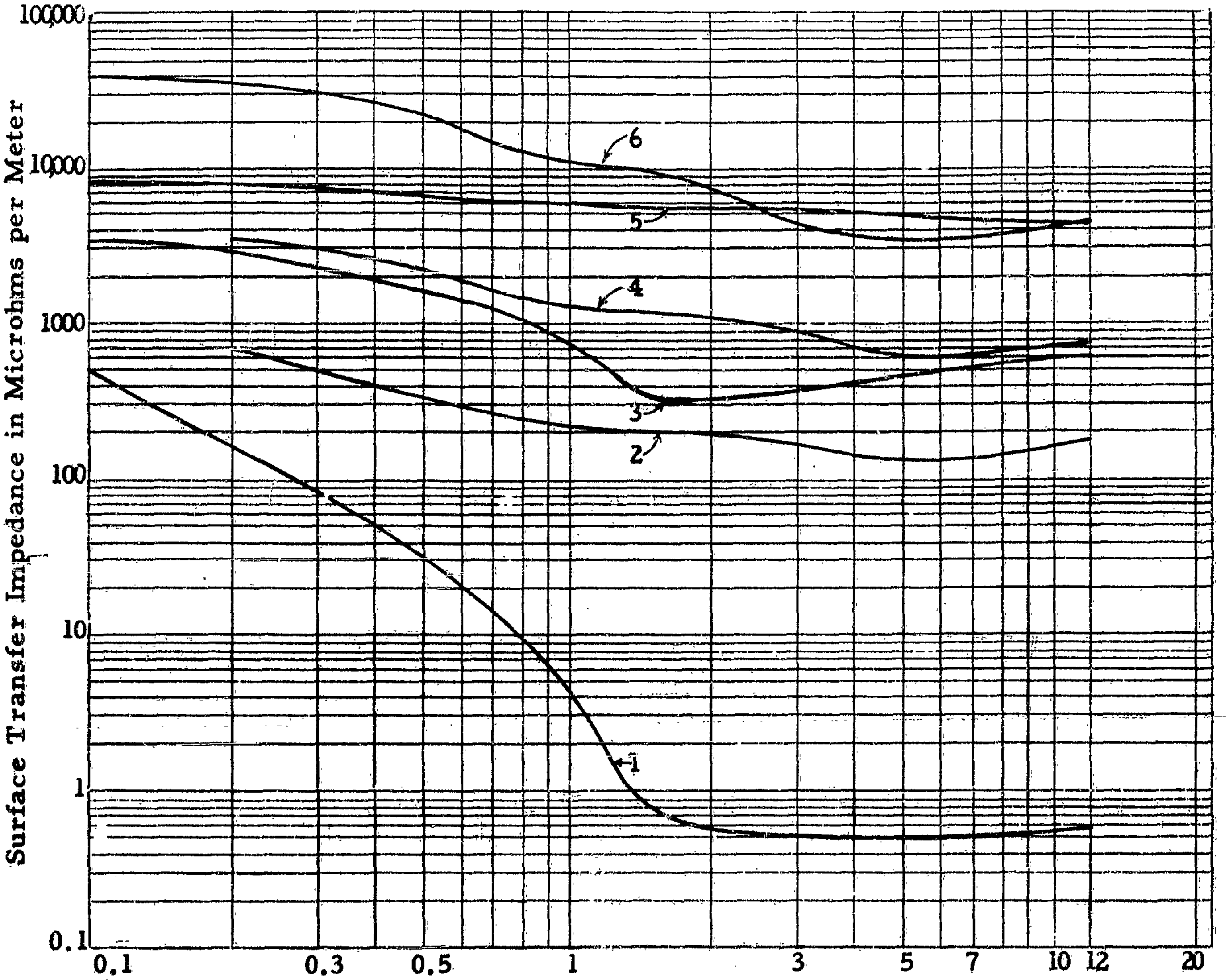


Fig. 3.1.2.11-C Shielding Effectiveness of Various Types of Flexible Conduit Described in Fig. 3.1.2.11-D

Specimen No.	Construction Details
1	Interlocked flexible metal hose, tinned stainless steel, plus two tinned-copper wire braids.
2	Interlocked flexible metal hose, tinned copper plus one tinned-copper wire braid.
3	Interlocked flexible metal hose, silver-laminated bronze, no braid.
4	Soldered convoluted brass or bronze hose plus one copper-clad steel wire braid.
5	Soldered convoluted brass or bronze hose plus one bronze-wire braid.
6	Square locked hose, aluminum plus aluminum braid.

Fig. 3.1.2.11-D Construction Details for Various Types of Flexible Conduit (See Fig. 3.1.2.11-C)

between adjacent convolutions. When a lesser degree of shielding suffices, a single braid may be used. If effective shielding at the lower frequencies is especially important, the use of magnetic material such as cold-rolled steel is recommended. In all cases, good electrical contact between adjacent convolutions is more important than low resistivity of the shielding material. As always, the maintenance problem is ever present. It is not sufficient that good electrical contact exist initially - it must be maintained under all service conditions.

As with rigid conduit, fittings and connectors pose special problems. These will be discussed in the next paragraph.

3.1.2.12 FITTINGS AND CONNECTORS

In fittings and connectors used to terminate either rigid or flexible conduit, the most important consideration, from a radio-interference point of view, is that of obtaining and maintaining continuous line contact between the mating members. The problems are the same as in the design of joints. Maintenance is usually the more difficult aspect of the two.

In joining the connector to the conduit, the metal tube or hose, as well as the covering braid or braids, if used, must be welded or brazed to the metal housing of the connector in order to insure good permanent electrical contact. Great attention must be given to mechanical strength, particularly with flexible conduits, because the points of maximum flexure are usually adjacent to the connectors.

The mating members of the connectors must be designed to provide a high pressure contact along a continuous line even under a slight misalignment of the shielding components. Tapered or wedge-shaped designs have been found satisfactory, and the conical, or spherical type shown in Figure 3.1.2.12, gives excellent results even under adverse service conditions. It should be emphasized that practically all connectors which rely for electrical contact entirely on the high pressure between the threads of AN connectors have been found unsatisfactory in service.

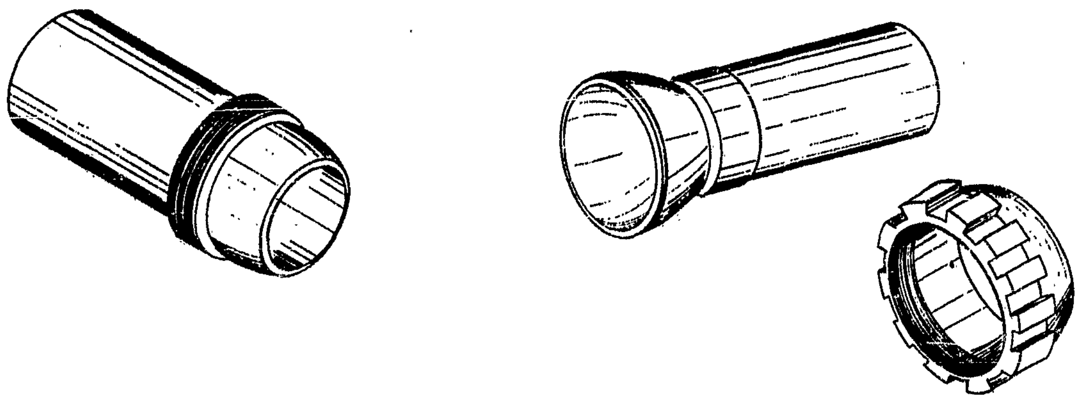


Figure 3.1.2.12 Disassembled Spherical-Type Connector for Flexible Shielding Conduit

3.1.2.13 ANTENNA RELAY SHIELDING

One of the most important examples of the application of shielding principles is the shielding of a receiver system. If a receiver has its own exclusive antenna, the receiver system consists only of the receiver proper, the antenna lead-in, and the antenna. But if the same antenna is shared by several receivers, or by a receiver and a transmitter, then the system must also contain a relay unit, which allows the antenna to be switched alternately to several components. Such antenna relays have often contributed to interference problems because they afford a coupling path to the receiver for interfering signals. In this paragraph, a general method is described which utilizes good shielding practices to overcome this difficulty.

The essential feature of the method is the housing of the entire antenna relay, together with any lightning-protection devices that may be necessary, in a shielded enclosure separate from both the transmitter and the receiver. Thus, an interference-free region is created within the aircraft which includes the receiver, the antenna lead-in, and the antenna-relay housing. This region is shown as unshaded in Figure 3.1.2.13-A, and the unprotected regions containing interference fields by cross-hatching. The transmitter and its antenna lead-in must be included in this unprotected region because the high voltages present at the antenna lead-in during transmission often prevent effective shielding methods at the transmitter.

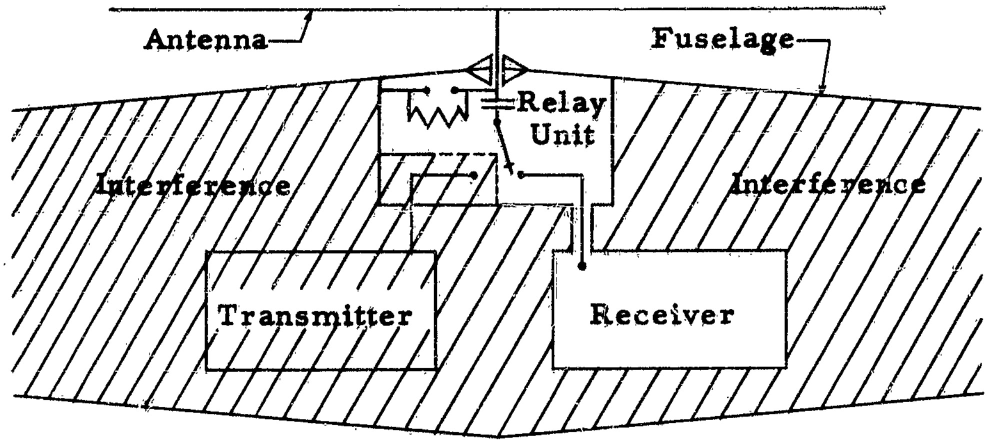


Fig. 3.1.2.13-A Schematic Diagram of Shielded Antenna System

The shielding conduit from the receiver to the relay-unit housing must usually be flexible because the receiver is shock-mounted. It is important to make the connections of this flexible cable absolutely "tight" at both ends. It is also necessary to keep the antenna away from openings in the skin of the aircraft in order to utilize the shielding properties of the skin of an all-metal aircraft.

Since the switch blade of the relay unit must make contact, at different times, both with the receiver and with the transmitter terminal, a special design must be used for the switching unit. Details of this design are shown in Figures 3.1.2.13-B and C. It is seen that the transmitter terminal is surrounded by a cylindrical shield with a longitudinal slot just wide enough for the switch blade to pass through. Such construction results in minimum impairment of the shielding effectiveness.

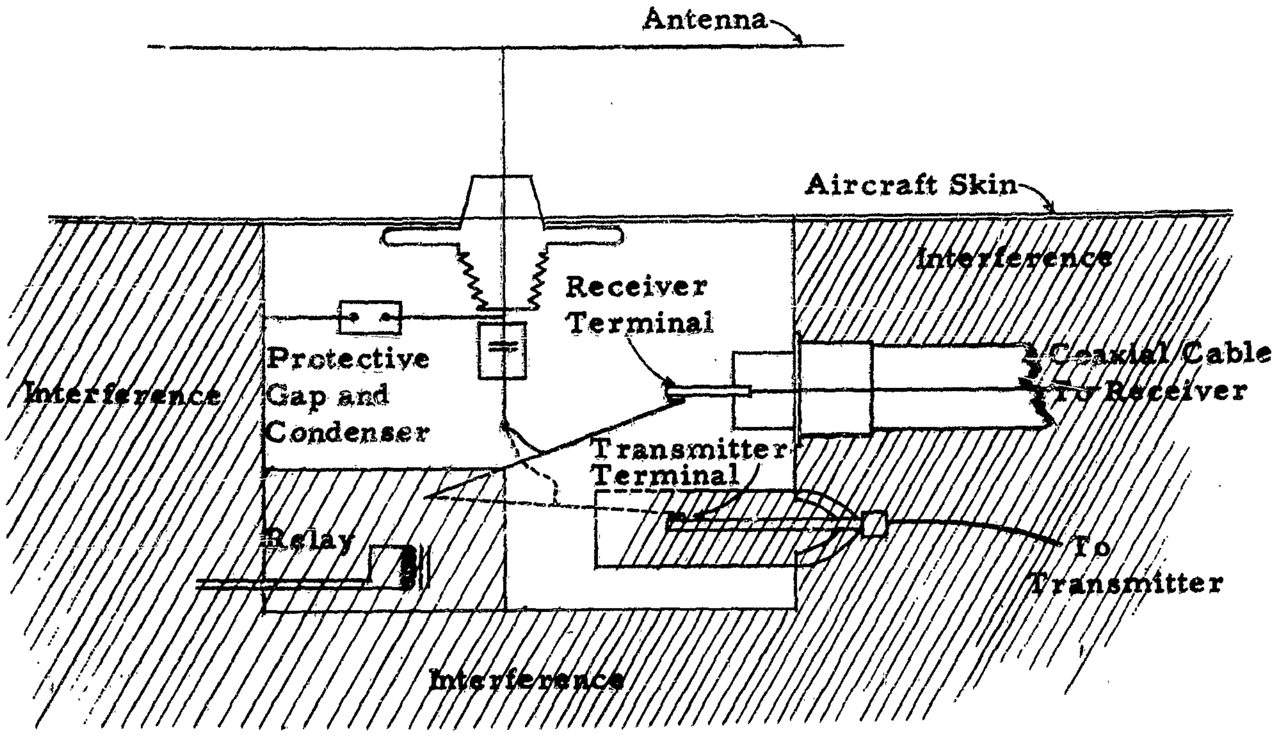


Fig. 3.1.2.13-B Details of Antenna Relay Shield

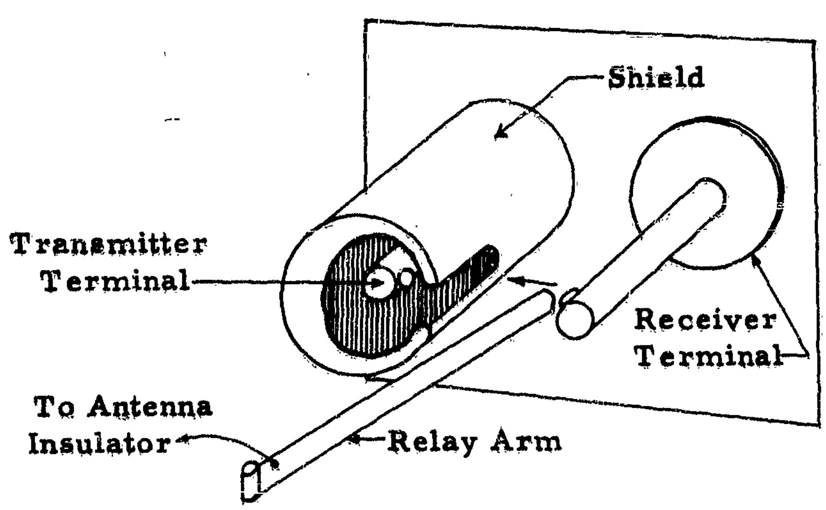


Fig. 3.1.2.13-C Transmitter-Post Shielding

3.1.3 BONDING

The purposes of bonding have been enumerated and explained in Paragraph 1.8.1.1. This paragraph deals with the techniques of bonding with special emphasis on those applications of bonding which have a direct or indirect bearing on the suppression of radio interference. As was pointed out before, poor bonding may contribute to radio interference in three distinct ways: It may allow electric charges to build up which can produce a spark, it may allow a varying electromagnetic field

to exist when the poorly bonded members are subject to shock or vibrations, and it may produce coupling paths due to the impedance of the poor bond being common to two circuits. The first and second conditions are usually found when structural parts of the aircraft are unbonded or poorly bonded. The third condition may exist in structural members also, but more often, it is found when electronic equipment, filters, or shields are inadequately grounded because of poor bonding.

Any one of the causes of trouble mentioned above is completely eliminated when a low impedance path is produced between the two conductors to be bonded. The impedance which is important in this connection is the impedance at radio frequencies. Actual measurements have shown that there is almost no correlation between the direct-current resistance of a bond and its radio-frequency impedance. Therefore, the direct-current resistance cannot be used as a measure of the effectiveness of bonding. A method of measuring the radio-frequency impedance of bonds is described in Appendix XII.

Moreover, even the measured radio-frequency impedance is not a perfect indication of the effectiveness of the bond in an actual installation. Since in the actual installation the artificial bond, a jumper, rivet, or screw connection, is always in parallel with some kind of natural bond, such as a direct metallic contact or the capacitance between two surfaces not in direct contact, the total impedance between the members to be bonded must be evaluated by considering the various parallel paths through which a radio-frequency current may flow. Thus, it may happen that a bonding jumper which was measured to have a satisfactorily low inductance combines with the capacitance of the system to form an anti-resonant circuit of extremely high impedance. This shows that bonding techniques must take into consideration the actual installation in which the bond is to be used.

In general, two main types of bonding may be distinguished: direct bonding and bonding by means of jumpers. Direct bonding is achieved by insuring permanent metal-to-metal contact between the members to be bonded. If this method is practical, it is always preferable. But when the contacts are subject to frequent separation, such as the edges of doors, or if clearance between the bonded members must be maintained for mechanical reasons, such as with control surfaces, or if the equipment to be bonded is shockmounted, then direct bonding is not feasible and bonding jumpers must be substituted. It must be borne in mind that a jumper is never more than a substitute for a direct bond. At best, its impedance may be only slightly larger than that of a direct bond. At worst, it may actually increase the impedance.

3.1.3.1 DIRECT BONDING

Direct bonding is accomplished by direct metal-to-metal contact between two surfaces under high and uniform unit pressure. <u>If properly constructed</u>, a bond of this type has a low ohmic resistance as well as a low radio-frequency impedance. Permanent joints of metallic parts made by welding, brazing, sweating, or swaging; semipermanent joints of machined metallic surfaces held together by lock-threaded devices, rivets, tie rods, or structural wires under heavy tension; pinned fittings driven tight and not subjected to wear; and, clamped fittings normally permanent and immovable: All these are considered as meeting the bonding requirements inherently <u>if all protective coatings are removed from the contact areas before assembly</u>. (See Military Specifications on Electrical Bonding for Aircraft.)

Bonds formed by direct metal-to-metal contact through mating surfaces held together by clamping devices may deteriorate with time. This is brought about by corrosive action which in time makes the bond ineffective by causing the contact resistance to increase beyond tolerable limits. Corrosive action may be either the galvanic or the electrolytic type or both depending on the nature of the metals in contact and on whether or not the metal-to-metal contact is part of a direct-current circuit; but, both types of corrosion take place only when moisture is in contact with the mating surfaces.

In the galvanic-type of corrosion, the two mating surfaces in contact with moisture act like a chemical cell of two metal electrodes immersed in a solution. In general, when an electrode is immersed in a solution, a potential difference develops across the junction of the electrode and the solution. The reason for this is that when a metal is placed in water, or any other ionizing solvent, some of the metal in contact with the water passes into solution as positively charged ions. This process leaves an equivalent amount of negative charge on the metal electrode. There is a tendency for such a process to occur at each of the two electrodes of the chemical cell which the moist mating surfaces of the bond resemble. However, if the electrodes are of the same metal and the solution in contact with them is homogeneous, no net potential difference across the two electrodes can be detected because the conditions at each are the same, exactly balancing one another. This would be the case for mating surfaces of the same metal exposed to moisture, and no such galvanic action would occur.

On the other hand, if the bond is formed by direct contact of dissimilar metals, in the presence of moisture, a chemical cell (resembling a voltaic pile or galvanic couple) is formed across which a net potential difference is developed. The reason for this is that each of the metals has a different tendency to go into solution as ions. Hence the potential difference (electrode potential) across the junction of one metal electrode and the solution is greater than that which exists across the other. Construction of a series of units, each made from two sheets of dissimilar metals separated by a wet cloth shows that metals can be arranged in an "electromotive series" so that each is positive when placed in contact with the one next below it in the series. This series, for metals commonly used in aircraft, is given in Figure 3.1.3.1-A.

Magnesium
Aluminum
Zinc
Chromium
Iron
Cadmium
Nickel
Tin
Lead
Copper
Silver

(Metals are listed in decreasing order of tendency to go into solution as ions)

Fig. 3.1.3.1-A Electromotive-Force Series for Metals Commonly Used in Aircraft

The chemical action* accompanying the establishment of these electrode potentials in a galvanic couple is such that the more positive electrode (higher in the series) corrodes by loss of metal while the other electrode does not. Hence, the nonreplaceable part of a joint formed by dissimilar metals should be a metal lower in the series than its mate. Moreover, the further apart any two metals are in the series means that a greater potential difference across the pair could be established and the chemical action (corrosion) would be more severe. Therefore, it is essential to select metals close to one another in the electromotive-force series when contact between dissimilar metals cannot be avoided. For example, the contact of a copper fitting and a magnesium casting would lead to excessive corrosive action because these two metals are too far apart in the electromotive-force series. This corrosive action could be minimized by plating the copper fitting with zinc or cadmium.

In addition to the galvanic action described above, there is another phenomenon, called electrolysis, which also produces corrosion due to chemical action. Electrolytic action takes place when a direct current flows between two metal surfaces in contact with a conducting solvent. The occurrence of electrolytic-type corrosion is independent of the nature of the metals in contact. It may occur along with galvanic action in joints of dissimilar metals, but, by itself, can account for the corrosion at joints formed by surfaces of the same metal in contact. Since the airplane structure and the casings of equipment are used for the ground return path of the direct-current power, there is the possibility of large DC currents through joints and connections serving as bonds to ground. If the joint contains moisture with dissolved salts or other impurities, the mechanism of electrolysis alone (by chemical reactions at the metal surfaces and the flow of ions) permits the passage of current, if no other path of much lower resistance exists. Depending upon the type of metal and the magnitude and direction of the voltage drop across the moist joint, dissolution and deposition of metal can occur at the metal surfaces of the joint. Moreover, chemical action of the dissolved impurities at the metal surfaces causes rapid contamination, corrosion, and destruction of the joint.

As explained above, the possibility of galvanic or electrolytic action necessitates the use of extreme care in assembling joints which serve as bonds for the ground return path. The surfaces should be absolutely dry before mating and held together under high pressure to minimize the chance of moisture creeping into the joint. After a joint is assembled with no moisture occluded, it is good practice to seal the periphery of the exposed edge with a suitable protective compound.

Direct bonding may be improved, and its use may be extended to surfaces in relative motion provided that the clearance between them remains very small at all times, by the use of non-hardening conductive silver pastes, for which the following claims are made by the manufacturer: They are rubbery, adhesive solids, resistant to oxidation and corrosion caused by moisture, heat or fumes. They exhibit the same electrical characteristics as solder, and can be used to replace it where physical contact between the parts is maintained by mechanical means. Furthermore,

—///—

*The above explanation of galvanic action is purposely simplified to give merely a qualitative picture of the phenomenon and to establish general rules for design practice. An exact description of the process, especially as to the actual seat of the EMF's developed, still lacks verification. Details concerning ion concentration and temperature dependence have been omitted from the discussion in this book.

if relative motion exists between the parts, a conductive paste which has the consistency of chewing gum has been designed to insure good electrical conductivity. Iron and other base metal contacts can be used in circuit-breaking equipment when treated with these pastes. Pastes are also effective in maintaining the electrical continuity of shields at seams, junctions, and joints.

The applications of the non-hardening silver conductive pastes as advocated by the manufacturer and summarized in the preceding paragraph must be justified since data on the conductivity of the pastes at radio and direct-current frequencies as well as data on their effectiveness in preventing oxidation and corrosion are not as yet available. Therefore, the conductive pastes must be subjected to extensive tests before any application is made, and caution should be observed when using them near moving parts.

3.1.3.2 BONDING JUMPERS

For direct or low-frequency alternating currents, bonding of equipment is easily accomplished. A wire or a length of tinned-copper braid suffices. However, at radio frequencies the same jumpers present considerable impedance. In order to understand the factors which determine the magnitude of the impedance, the equivalent circuit shown in Figure 3.1.3.2-A must be analyzed. In this diagram, R is the ohmic resistance including the increase due to the skin effect, L is the total series inductance of the jumper, and C is the combined capacitance due to the distributed capacitance of the jumper and the capacitance of the bonded members.

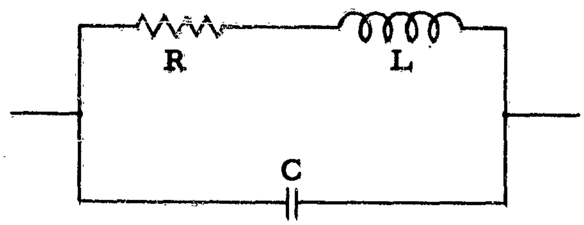


Fig. 3.1.3.2-A Schematic Diagram of a Bonding Jumper

Experiments have shown that the effective value of the resistance despite its rise due to skin effect at higher frequencies is negligible except near the point of anti-resonance. If the resistance is neglected, formulas for the impedance Z of the equivalent circuit are given as:

$$|Z| = \frac{\omega L}{1 - \omega^2 LC} = \frac{1}{\omega C}\left[\frac{1}{1 - \frac{1}{\omega^2 LC}}\right] \qquad (3\text{-}28)$$

where ω is the angular frequency.

If $\omega^2 LC$ is less than one, the circuit operates below its anti-resonant frequency and acts as an inductance. An increase of the values of capacitance or inductance, up to the point where $\omega^2 LC$ equals one, results in a decrease of the impedance. If the circuit operates above the anti-resonant point, $\omega^2 LC$ is greater than one, and the circuit acts as a capacitance. With an increase of the values of capacitance or

inductance the impedance decreases. Therefore, to keep the radio-frequency impedance low at frequencies below the anti-resonant frequency, the values of capacitance and inductance must be comparatively low, but for frequencies above the anti-resonant frequency, the values of capacitance and inductance must be comparatively high to obtain low values of impedance as shown in Figure 3.1.3.2-B, C, and D.

The region of frequencies of interest is almost always such that the anti-resonant frequency of the jumper occurs near the upper end of this region. In order to obtain values of radio-frequency impedance as low as possible, it is necessary to have a comparatively low LC product. Lowering the LC product raises the anti-resonant frequency, and, as may be seen from Figure 3.1.3.2-E, lowers the impedance in the range of the frequencies of interest.

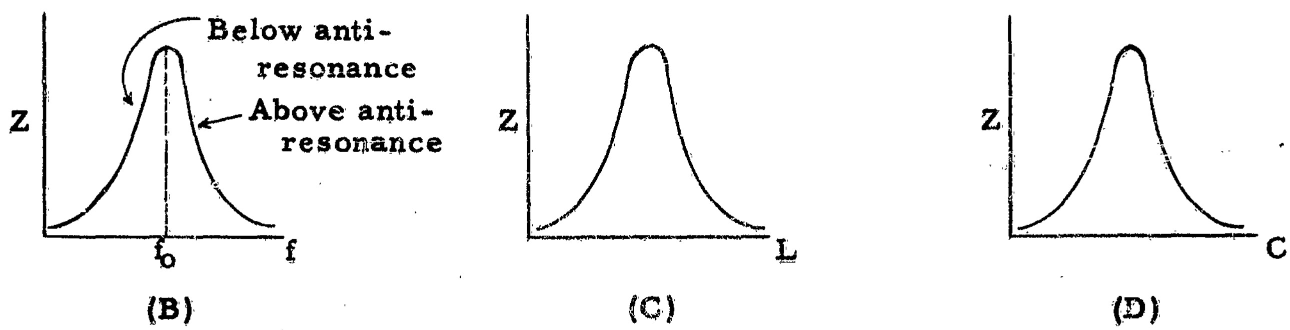


Fig. 3.1.3.2-B, C, D Magnitude of Impedance of a Parallel Circuit As a Function of: (B) Frequency, (C) Inductance, and (D) Capacity

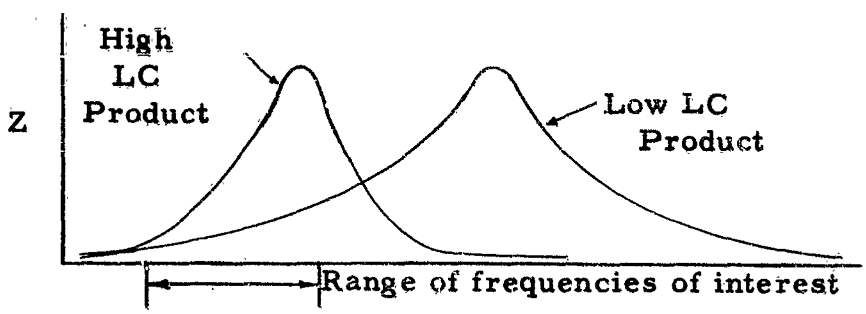


Fig. 3.1.3.2-E Magnitude of Impedance of a Parallel Circuit As a Function of Frequency for Two Different Values of the LC Product

Experiments have shown that the physical characteristics of a jumper have a marked effect on their radio-frequency impedance. An increase in the length of a jumper causes its impedance to increase proportionally, but an increase in either its cross-sectional or its surface area causes a non-proportional decrease in its impedance. However, the change in the surface area exerts a greater effect on the impedance of a jumper than a corresponding change in its cross-sectional area. In the design of a bonding jumper, these results must be considered along with the requirements noted in Appendix XV.

For certain radio-frequency-suppression bonding applications, "round" bonding jumpers consisting of strands of wire arranged in a rope twist can be substituted for the more expensive "flat-braid" bonding jumpers consisting of woven strands of wire. Radio-frequency measurements made in the range of 0.15 to 30 megacycles show that the impedance of the round jumper is only slightly higher than that of the flat braid of comparable size at all frequencies within this range. The curves of Figure 3.1.3.2-F indicate this clearly. They also show the direct, approximately linear, relationship between the impedance and the frequency. This is evidence that the resistive component of the radio-frequency impedance is comparatively low, and that its reactive component is inductive. Despite the slightly higher radio-frequency impedance values of the round bonding jumpers, their use is justified in view of the greater dependence of impedance upon other conditions and characteristics such as the length of the bond and its orientation with respect to the ground plane.

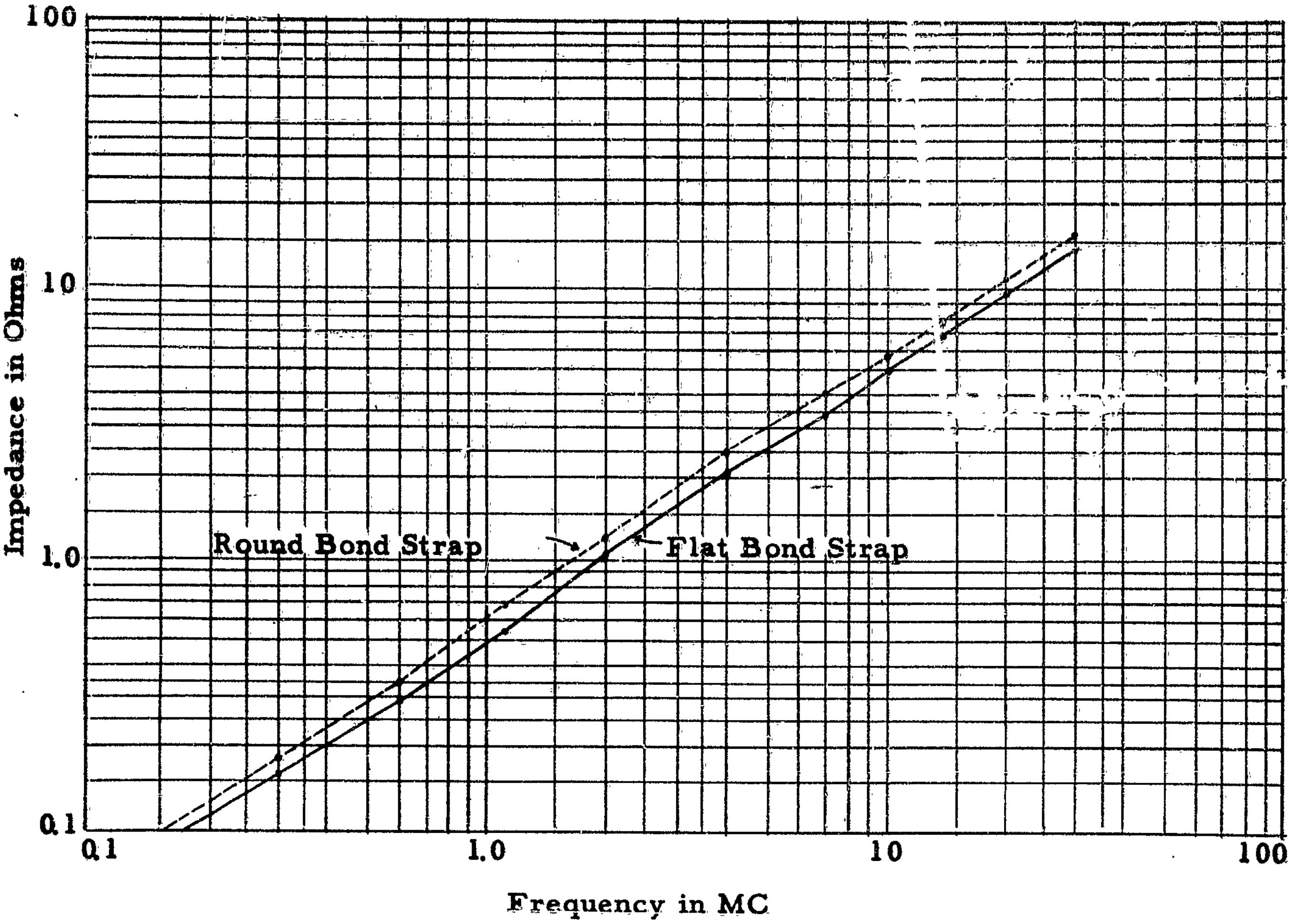


Fig. 3.1.3.2-F Impedance Characteristics of Flat and Round Bonding Jumper

In Paragraph 1.8.1.1 several applications of bonding other than the suppression of radio interference are given. Among these is the minimization of lightning damage. To provide adequate bonding for lightning protection, bonding jumpers of tinned-copper stranded cable should have a minimum cross-sectional area of 6475 circular mils to withstand a maximum current surge of 100,000 amperes built up in 10 microseconds and damped to one-half its maximum value in 20 microseconds. If stranded aluminum cable is used, its minimum cross-sectional area should be 10,000 circular

mils. Twisted cables are superior to braided cables of the same cross-sectional area because the latter crystallize and break more readily than the former due to the magnetic stresses accompanying the surge currents. Due to the oscillatory nature of the discharge, stranded wire must be used to minimize the skin effect.

Although the effective resistance provides a fairly accurate evaluation of the effectiveness of bonding when applied to lightning protection, it does not have, as previously mentioned, any significance in indicating the effectiveness of bonding used to suppress radio interference. As far as the bonding of external structural parts is concerned, lightning protection is the most important consideration, and it must be assumed, due to the lack of additional data, that a low effective-resistance bond, adequate for lightning protection, is also adequate for the suppression of radio interference.

Bonding jumpers for shockmounted equipment pose special problems because of the limited amount of space available for them, because they form part of the mechanical system in addition to serving as carriers of currents, and because good grounding is especially important for such shockmounted equipment as receivers. The development of special bonding jumpers for shockmounted equipment is described in detail in Appendix XV.

3.1.3.3 BONDING OF STRUCTURAL PARTS

The discharge of static charges that accumulate on the surface of an airplane is a source of radio interference (see Section 4). The effects of these discharges may be kept to a minimum by effective bonding of all the elements in the vicinity of the antennas, and by the installation of static-dischargers, described in Paragraph 4.3.8.

The engine cowling consists of several sections - some removable for ease of maintenance and servicing - each of which could accumulate static charges, the discharge of which would create interference that is easily picked-up on nearby antennas. Therefore, each section of the cowling that does not make good electrical contact to the basic structure should be bonded to the structure by a braided bonding jumper of about 1/8-inch inside diameter. Round braid should be used instead of flat braid in order to reduce breakage.

To secure a good bond between the engine and the engine-mount structure, at least four braided bonding jumpers of about 3/4-inch inside diameter should be employed. If shock-absorbing jumpers are used at the fire wall, similar jumpers should be installed across each unit.

3.1.3.4 BONDING OF TUBING AND CONDUIT

The outer surface of long spans of conduit or tubing is a possible high-impedance path for interfering currents from sources outside the tubing or conduit. To minimize this possibility, such spans should be properly bonded to ground at both ends and several intermediate points.

Ordinary clamps cannot be used to bond flexible conduit because the pressure exerted on a comparatively small surface area of the conduit is sufficiently high to

compress it or force it to give way. To overcome this a flared split-sleeve is fitted around the conduit, as shown in Figure 3.1.3.4-A, which distributes the high pressure delivered by the bonding clamp over a larger area, thereby resulting in a low unit pressure on the conduit. Contact is further improved by soldering the sleeve to the conduit, when the materials permit, through several holes in the sleeve provided for this purpose.

Figure 3.1.3.4-B illustrates a method for bonding rigid conduit or tubing to a structure through supporting attachments. The number of mechanical supports required is generally adequate to provide an efficient bond even when the conduit is carrying interference signals.

The conduit or tubing to which bonding clamps are attached should be cleansed of paint and foreign material over the entire area covered by the clamp. All insulating finishes should be removed from the contact area before assembly, and anodized screws, nuts, or washers should not be used for attaching parts in making bonding contact. If, in bolting the bonding clamp to the structural surface, a star washer is used, as shown in Figure 3.1.3.4-B, any protective coating (unless very thick or tough) need not be removed from this surface since the points of the washer penetrate to the bare metal.

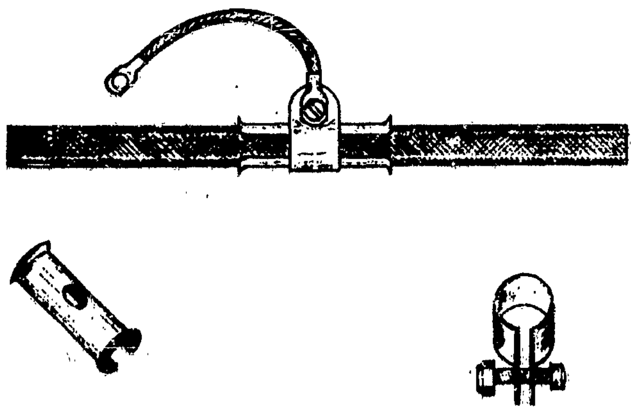


Fig. 3.1.3.4-A An Acceptable Method of Bonding Flexible Conduit

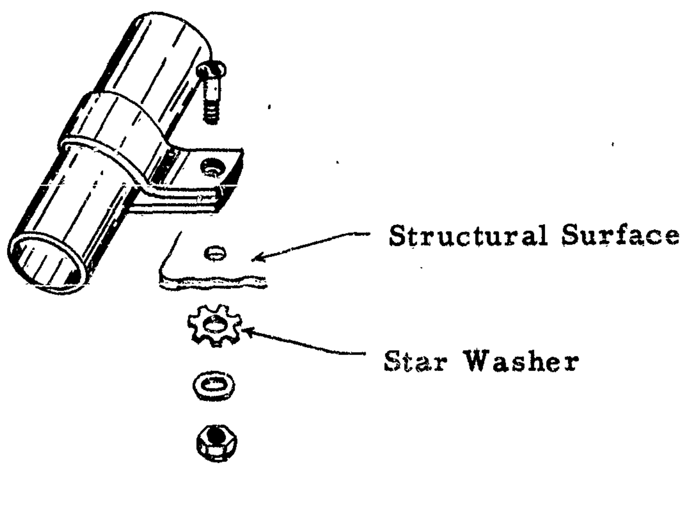


Fig. 3.1.3.4-B Tubing Clamp Bonded to Painted Surface

3.1.4 SPECIAL CIRCUITS

In Paragraph 1.8.3.3 the basic types of circuits used for interference reduction in receivers were enumerated. Specific circuits, which have proven useful in practice, are given here and their operation is explained briefly. Examples are given of limiters and wave traps, which find extensive practical applications with a wide variety of design depending on their purpose. Limiters are most useful if the interference is of the impulsive type, i.e., if it consists of pulses of large amplitude and very short duration. Parallel-tuned traps are used when disturbances of known frequency enter a receiver through the medium of the antenna.

3.1.4.1 LIMITERS

When designing limiters as a means of interference suppression, it must be borne in mind that their effectiveness is highest when the frequency selectivity in the circuits after the limiter is greater than the selectivity preceding the limiting.

The limiter shown in Figure 3.1.4.1-A operates satisfactorily on both modulated and unmodulated reception. It is a simple, convenient type requiring only a fixed capacitor, two fixed resistors, and an independent diode besides the normal components of a diode second detector.

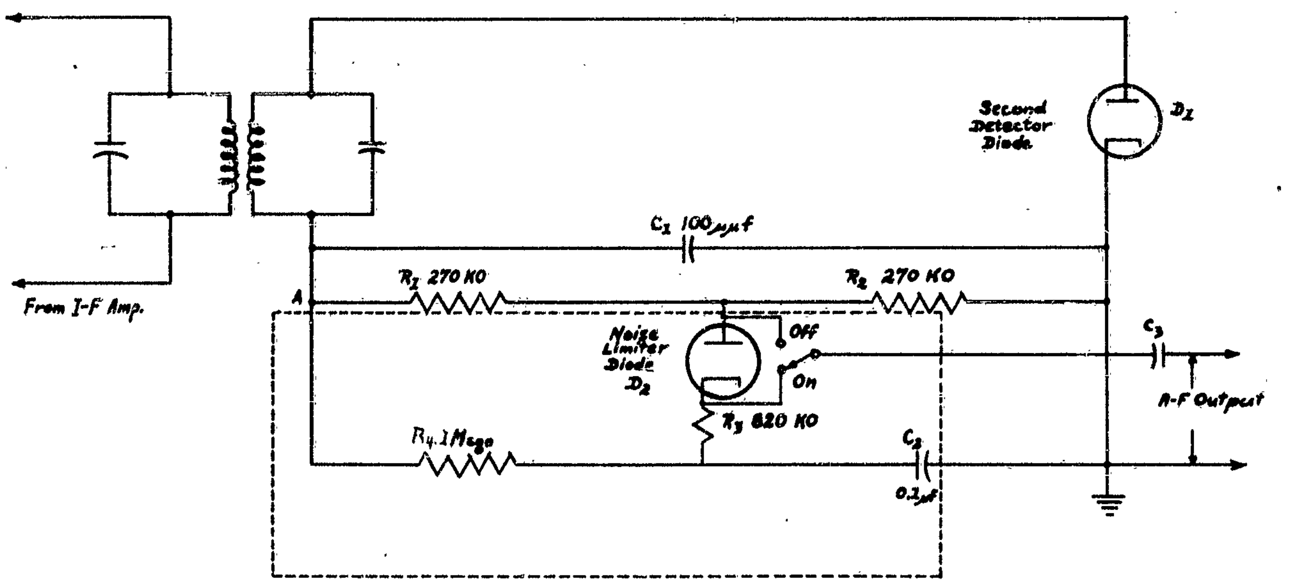


Fig. 3.1.4.1-A Series Limiter

When the switch is in the "OFF" position, C_3 connects to junction point B of the detector diode load resistors, R_1 and R_2, and the limiter diode has no effect on circuit performance. When the switch is "ON", C_3 is connected directly to the cathode of D_2, putting the diode switch in the circuit.

Assuming a potential of 10 volts across R_1 and R_2 by a constant carrier, the cathode of the limiter would be 10 volts negative with respect to ground if the diode were not conducting. The plate, connected to point B, is 5 volts negative with respect to ground. Hence the plate is 5 volts positive with respect to the cathode, and the

limiter diode conducts, having fairly low resistance compared to other circuit resistances. The output capacitor, C_3, is then connected to point B through the limiter diode, providing audio frequency output. This output is reduced to 45 percent of what it would be without the limiter, but generally the reduction is of little significance.

The difference in time constants between plate and cathode circuits allows the diode resistance to become very high when a large interference voltage appears, effectively preventing conduction through the interference limiting diode, D_2, and cutting off C_3 from point B. The amplifier will have no appreciable input for the duration of the interfering signal. By the time the cathode of the limiter diode goes negative with respect to its plate, the interfering signal will have decayed, restoring audio frequency input to the amplifier.

Distortion in this limiter is noticeable on an oscilloscope only above 40 percent modulation. Speech and coded transmissions maintaining an average modulation level of 30 to 40 percent are commonly encountered.

The modified shunt type of noise-peak limiter circuit, Figure 3.1.4.1-B, is similar to the series type except that the plate of the limiter diode and the low end of the cathode resistor are interchanged. When an interference peak makes the diode D_2 conduct, it acts to reduce the output voltage. Grounding the low end of the intermediate frequency secondary increases the stability of the system.

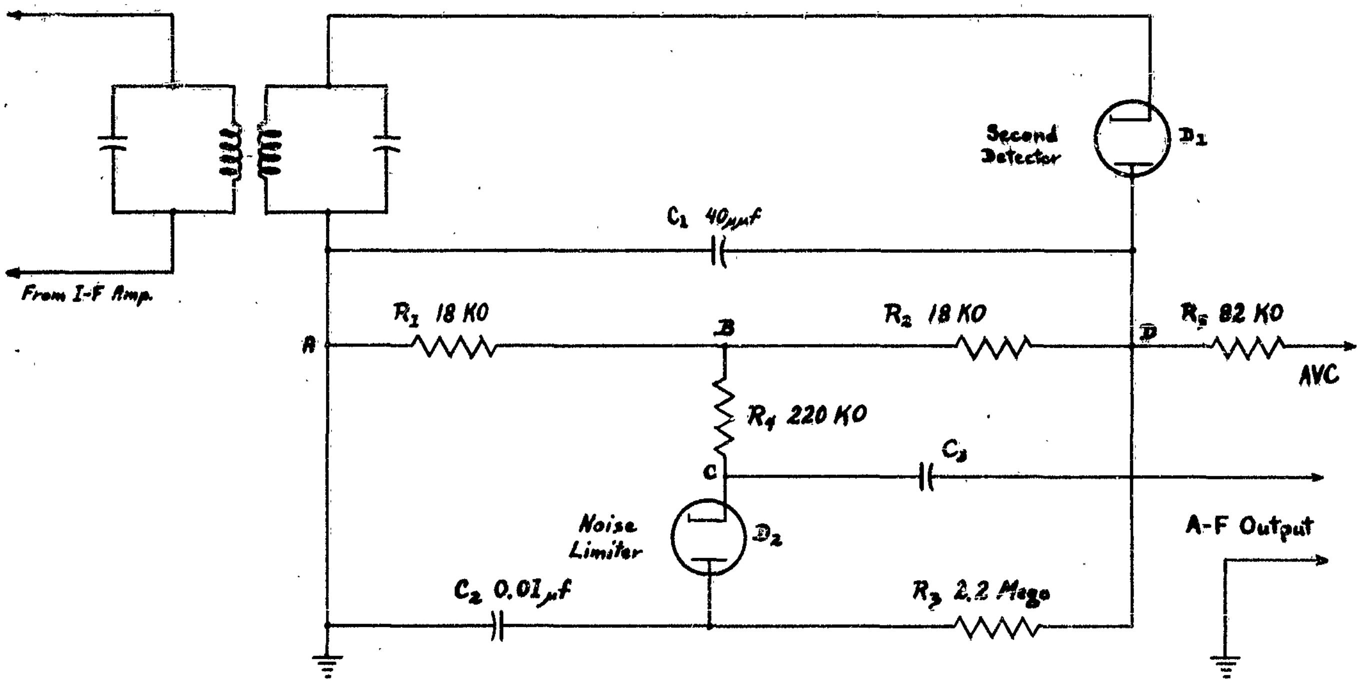


Fig. 3.1.4.1-B Shunt Limiter

In this circuit normally the cathode is positive with respect to the plate, and D_2 is not conducting. The time constant in the plate circuit is more than 10,000 times longer than that of the cathode circuit, so that any interference pulse in excess of normal bias drives the cathode negative and the diode conducts, shunting the input

of the following audio frequency stage. Shunting action is more complete with the addition of R_4 which acts as part of a voltage divider when D_2 is conducting and helps attenuate detector-load voltage peaks.

The limiting action ceases when the interference pulse decays or C_2 charges, since then point C becomes positive with respect to point D. Audio distortion begins at about 100 percent modulation for values shown. If the ratio of R_2 to R_1 is 0.4, distortion begins at approximately 40 percent modulation. This limiter is not as good as the simple series type at lower carrier frequencies.

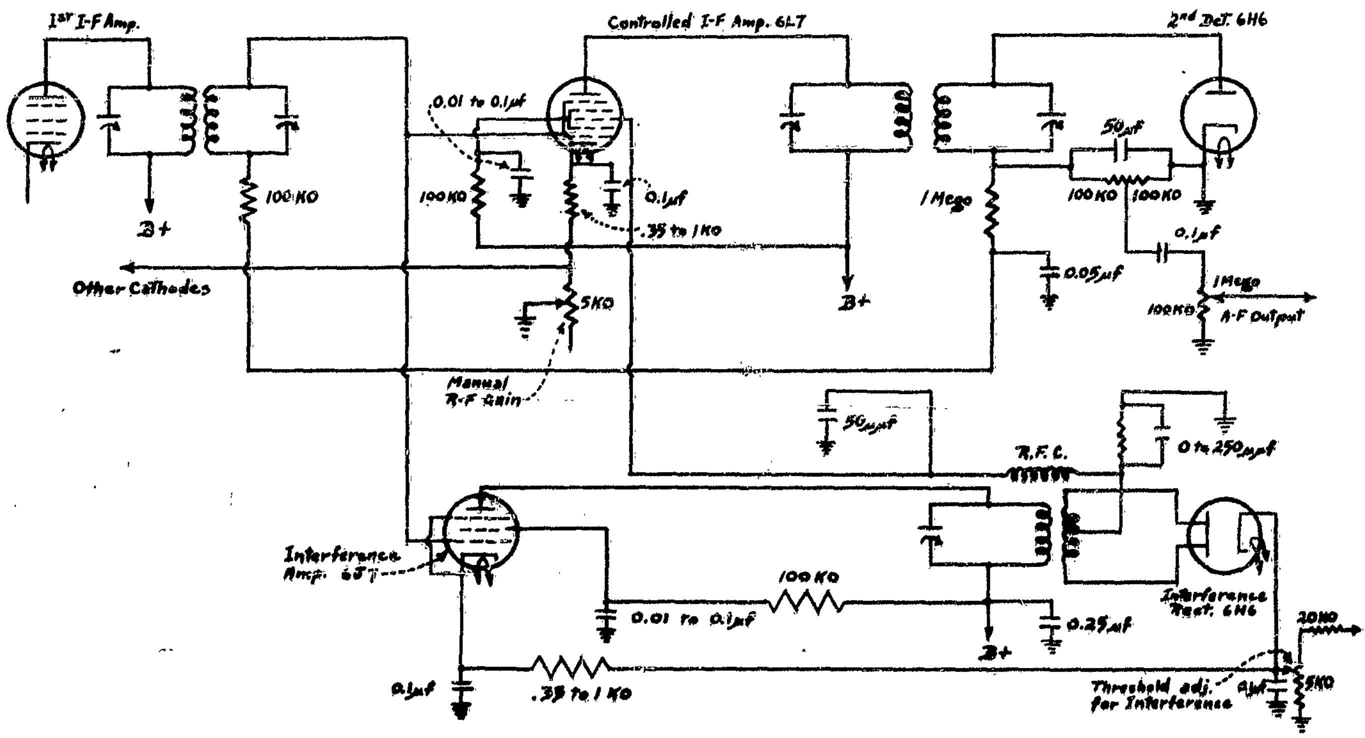


Fig. 3.1.4.1-C Combined Limiter-Blanking Circuit

The system given in Figure 3.1.4.1-C for interference suppression silences the receiver momentarily when an interfering pulse stronger than the desired signal is received. Here the blanking and limiting principles are combined.

The interference amplifier (6J7) has its grid connected in parallel with the grid of the final intermediate frequency amplifier, and delivers its output to an auxiliary rectifier. The direct-current bias thus obtained is applied to the third grid of the 6L7 final intermediate-frequency amplifier. When properly adjusted any interference voltage whose peak amplitude exceeds the signal being received will develop enough bias to make the final intermediate frequency amplifier tube inoperative, thus silencing the receiver for the duration of the pulse. This is accompanied by some distortion, but reception is much improved over that obtained without the interference suppressor.

The same effective principle is applied in the counter-modulation type of interference-reducing circuit shown in Figure 3.1.4.1-D.

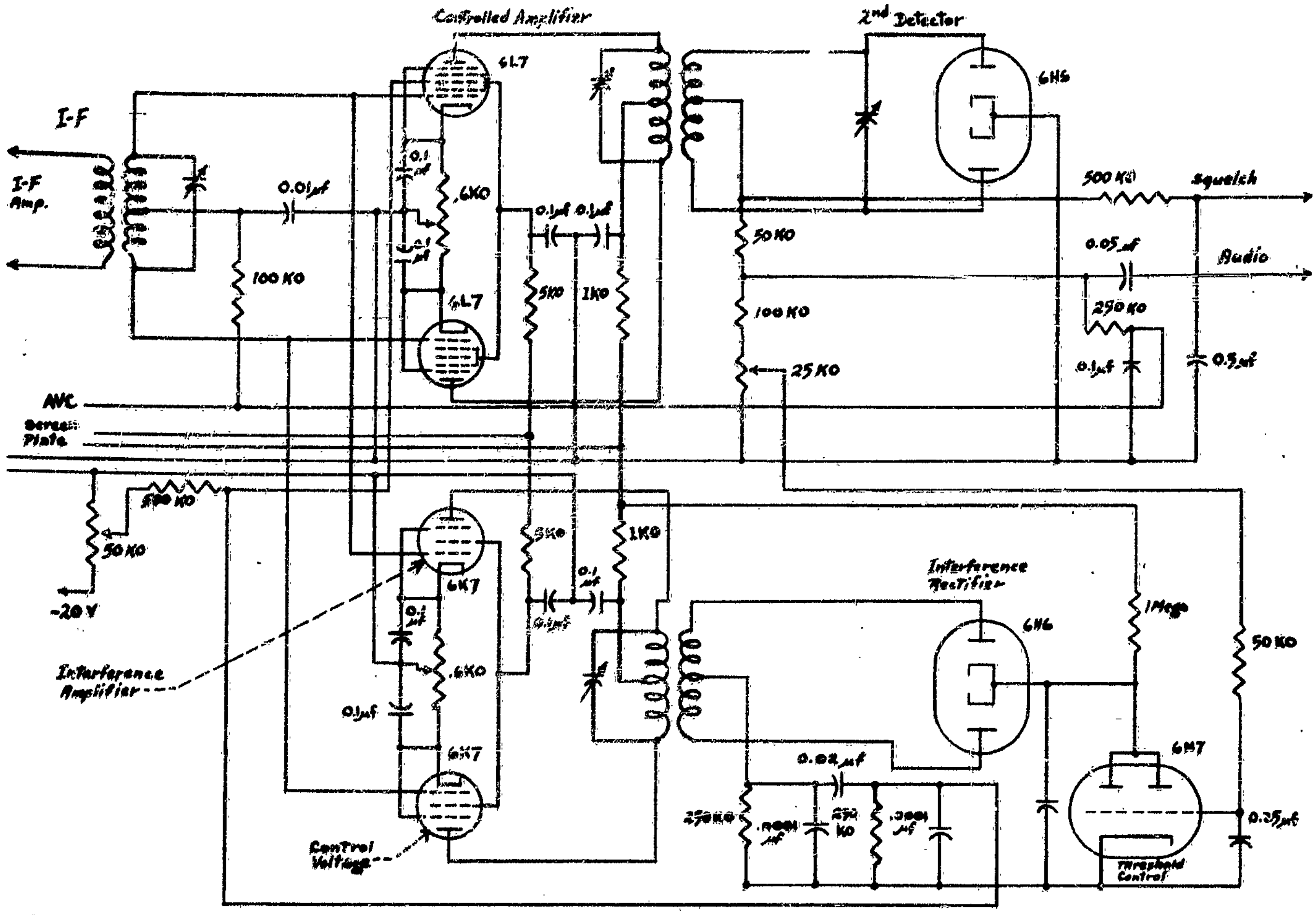
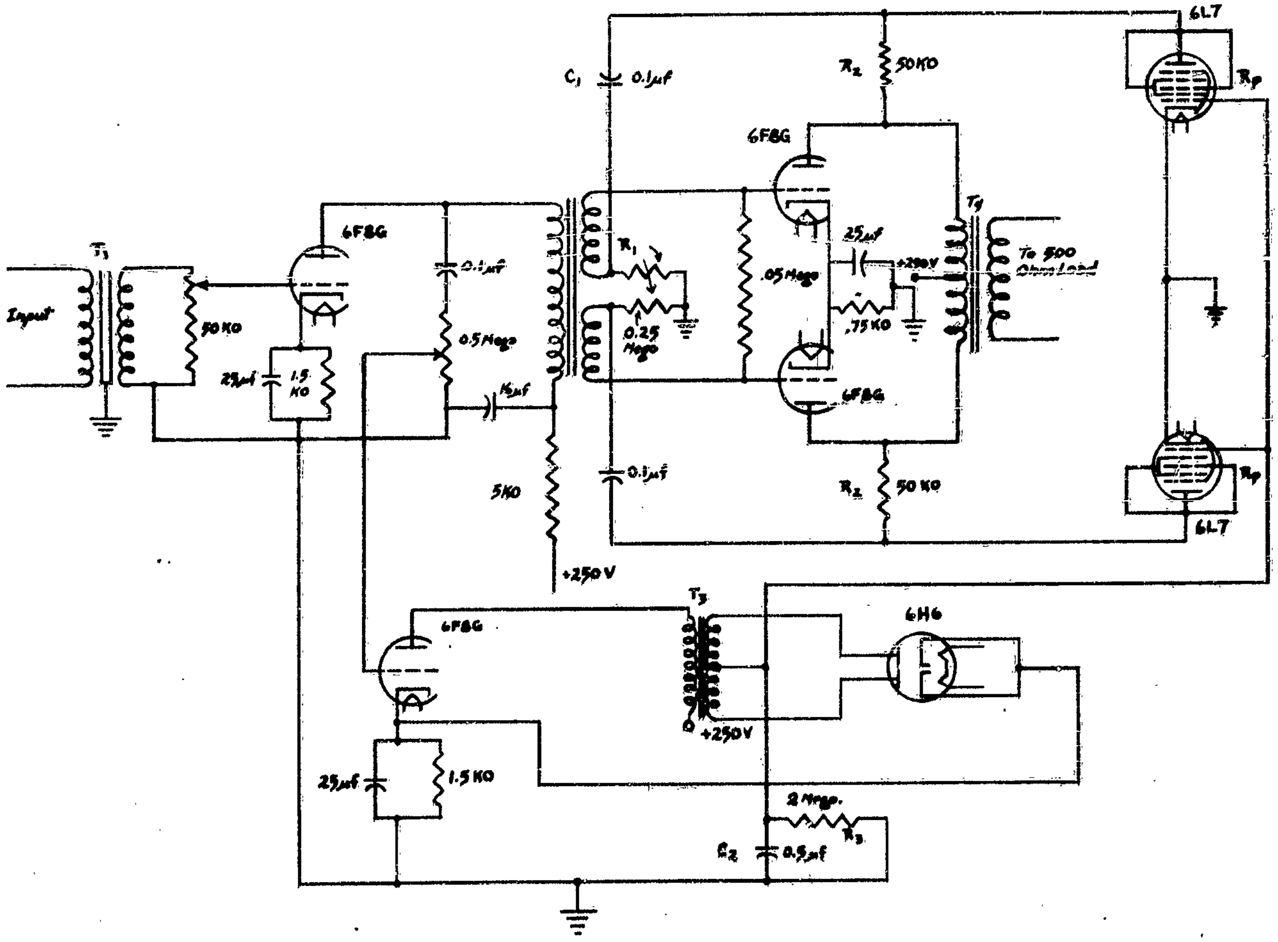


Fig. 3.1.4.1-D Limiter Employing Counter-Modulation

In this circuit, intermediate frequency signals and interference voltages are injected into a push-pull controlled amplifier stage, paralleled by a push-pull interference amplifier whose plate circuit is tuned lower than the intermediate frequency value. This allows more interference and less signal voltage at the interference-rectifier input than at the second detector input. Broad tuning of the intermediate frequency stages permits sufficient detuning of the interference amplifier to virtually eliminate signal voltage in this circuit. Any signal voltage still present is prevented from affecting the circuit operation by the automatic threshold control tube. The threshold control tube regulates a variable-delay bias for the interference rectifier, allowing rectification only above the level of the signal component in the voltage being rectified. The automatic volume control voltage fed to the grids of the threshold tube regulates the delay level for a change of signal strength, and the interference component is taken from the interference rectifier load resistor. Then it is fed through a blocking condenser to the 6L7 grids. The level of zero axis for the counter-modulating voltage is set by a direct current bias supplied to these grids. Once the delay and bias adjustments are made, they need not be changed.

Though not especially simple, the circuit is effective and fully automatic. It reduces the interference voltage before rectification and does not increase the automatic volume control voltages. Thus receiver gain, interference rectifier delay,

and interference amplifier gain are all controlled by the strength of the desired signal.



3.1.4.1-E Limiter Employing Feedback

The limiter circuit shown in Figure 3.1.4.1-E employs degeneration as a means of obtaining limiting action. The input stage of the amplifier uses one triode section of the 6F8G tube, the other section being used for an auxiliary amplifier for the 6H6 control rectifier. The input stage is transformer coupled to a push-pull output stage using another 6F8G tube. Degeneration is used on this stage, and the feedback factor is determined by R_1, R_2, and the plate resistances of two 6L7 tubes. The feedback factor can be controlled by varying the plate resistances of the 6L7 tubes since they are effectively in parallel with the resistors R_1. An increasing signal causes the grid bias on the 6L7 tubes to increase, which increases the plate resistance and the feedback factor and results in decreased gain, thus producing compression. Negative feedback has the added advantage of holding distortion to a low value.

By putting an initial positive bias on the 6H6 full wave rectifier, the control of the auxiliary gain will delay compression till any desired output is reached, within the limitations of the amplifier. An amplifier plate resistance at 7700 ohms, a transformer ratio from primary to secondary of 2 to 1, and a resistance of 1000 ohms for the 6H6, in conjunction with 0.5 microfarads for C_2, give an acting time of 1.5 milliseconds. Releasing time, with R_3 equal to 2 megohms, is 1 second.

A push-pull compressor stage with balanced feedback network is necessary to eliminate transient distortion due to compressor action, and is also desirable from the standpoint of low inherent distortion. Noise and hum level is 75 db below 6 milliwatts output, largely because of a regulated power supply.

In broadcast service, this amplifier is able to handle all ordinary peaks. There are no thumps when compression takes hold, and the general operation is quite smooth. Feedback does not help reduce interference at low power levels, such as thermal agitation, induced hum voltages, and microphonics.

Where automatic volume control is employed in a receiver of considerable sensitivity, a disagreeable amount of interference will be heard in the output when no carrier is present. An arrangement for blanking the receiver during the tuning process, often referred to as a squelch system, will suppress this form of interference and is indicated in typical schematic form in Figure 3.1.4.1-F.

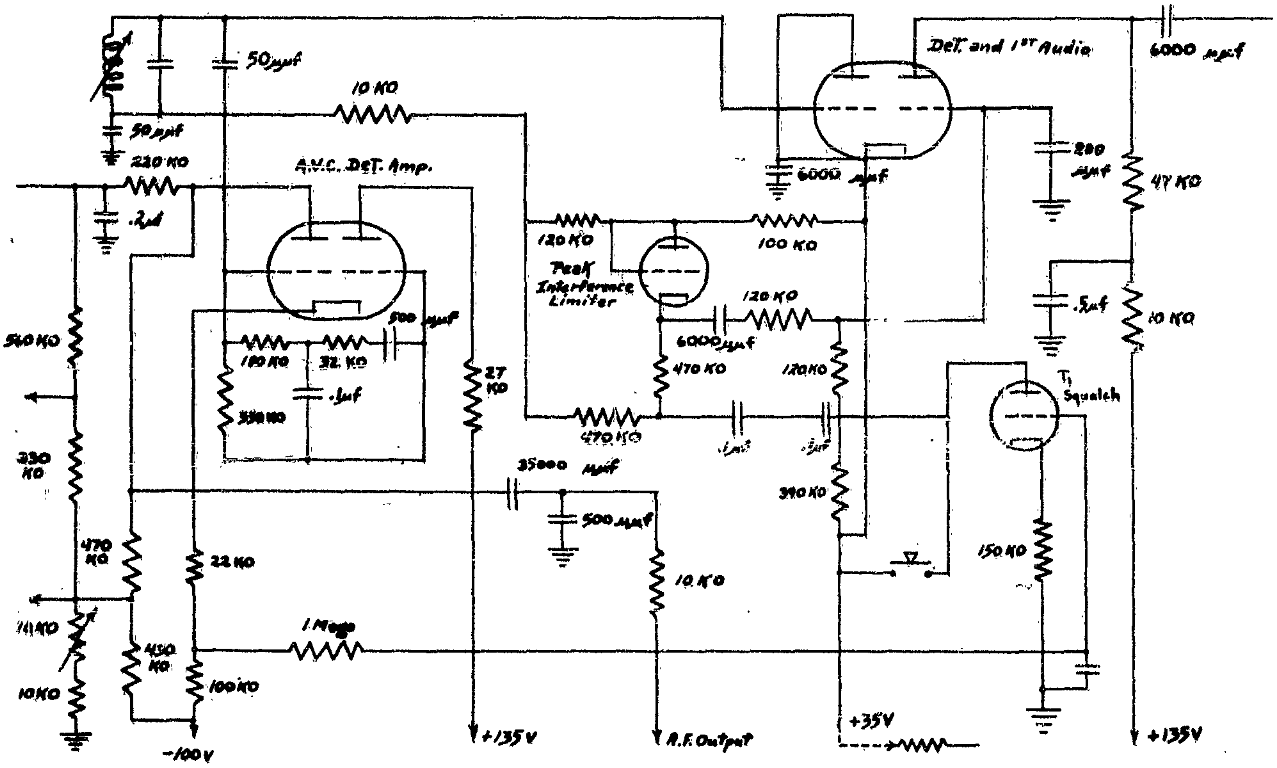


Fig. 3.1.4.1-F Squelch Circuit

The first audio tube is biased beyond cut-off by the action of Tube T_1 unless the grid bias of Tube T_1 approaches or exceeds cut-off. By using the automatic volume control system to bias T_1, the receiver can be made inoperative until a carrier of pre-determined amplitude is present.

A modification of this arrangement is to operate Tube T_1 from a separate branch of the intermediate-frequency amplifier that delivers its output to a second diode. By making this auxiliary intermediate-frequency branch very selective, the signal-to-interference ratio will be much higher for the interference-suppressor diode than for the detector diode. The threshold level of the system may be set such that the signal is so low with respect to the interference as to be barely usable. The receiver will then deliver no output until tuned exactly to the desired carrier frequency.

3.1.4.2 WAVE TRAPS

A wave trap in its simplest form, as explained in Paragraph 1.8.3.3.2, consists of a parallel-tuned circuit connected in series with the antenna of a receiver. It should be designed to resonate at the frequency of the interfering signal. The circuit of a simple trap is shown in Figure 3.1.4.2-A.

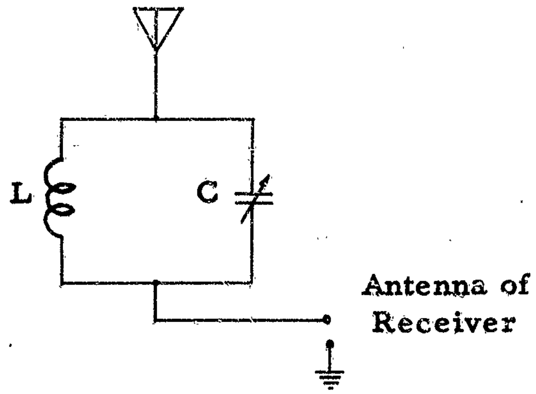


Fig. 3.1.4.2-A Basic Wave Trap

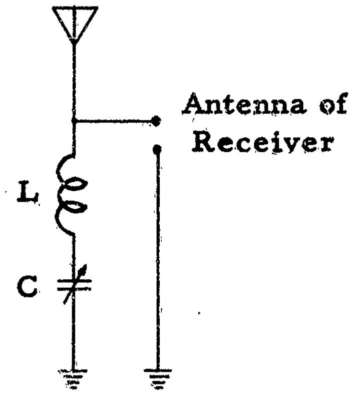


Fig. 3.1.4.2-B Series-Resonant Wave Trap

Traps are readily designed to offer a high impedance to an undesired frequency or band of frequencies. A high C/L ratio and high Q are most desirable for proper attenuation of the interference as well as low insertion loss at the desired frequencies. They are particularly useful in rejecting an interfering frequency at or near the intermediate frequency of the receiver. In this case the best present designs have an insertion loss of the order of 1 db with a maximum of 2 db.

A wave trap may also consist of a series resonant circuit in parallel with the receiver as shown in Figure 3.1.4.2-B. In this case a high Q and a low C/L ratio are desirable.

Suitable values for the elements of the circuit of Figure 3.1.4.2-A, for selected frequencies, are given in Figure 3.1.4.2-C. Wave traps consisting of lumped elements of this type are rarely used for frequencies above 30 to 40 megacycles.

Frequency in Megacycles	Capacitance in μμf	Inductance in μ	Coil Design Data
3.5	140	16	32 turns No. 22, 1"dia., 1"long
7	100	6	19 turns No. 22, 1"dia., 1"long
14	50	3.5	14 turns No. 18, 1"dia., 1"long
21	35	2.2	12 turns No. 18, 1"dia., 1"long
28	25	1.5	9 turns No. 18, 1"dia., 1"long

Fig. 3.1.4.2-C Representative Values for Basic Wave Trap

In several types of aircraft, the following types of wave traps have been successfully applied in reducing interference from IFF and similar systems operating

at frequencies about and above 50 megacycles.

(a) A radio frequency choke coil, designed to resonate with its distributed capacity, is installed in the antenna circuit of the communication receiver. Such a coil consists of three series windings of fifteen to thirty turns each, depending on the interfering frequency, wound on a bakelite form roughly 3/8 inches in diameter and 4 inches long.

(b) A quarter-wave-length open-circuited stub is connected between the antenna post and ground of the communication receiver. This stub consists of No. 18 solid copper, insulated, push-back type wires, about one quarter of a wavelength long and twisted about 9/10 of their length, the remainder serving as leads, as shown in Figure 3.1.4.2-D. The stub should be cut back experimentally until maximum interference reduction is obtained. This point is critical, and the wires should not be cut back more than 1/8 inch at a time in order to make sure that this point is not missed. A short length of coaxial cable may be used instead of the twisted wires. Its length must be adjusted by trial and error in the same way as described above for the twisted wires.

(c) A three-section parallel-resonant series wave trap constructed from coaxial cable, similar to the radio frequency choke coil, may be inserted between the antenna relay and the receiver antenna post. Each section is approximately a quarter wave length long and short-circuited, as shown in Figure 3.1.4.2-E. The exact dimensions are determined by trial and error.

The choke coil described in (a) is suitable in the range from about 40 to 100 megacycles. Types (b) and (c) are suitable above 100 megacycles.

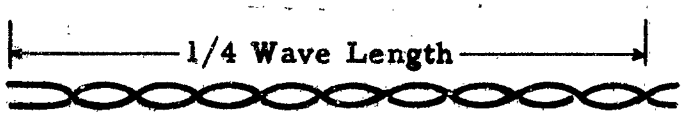


Fig. 3.1.4.2-D Twisted Quarter-Wave Series-Resonant Parallel Wave Trap

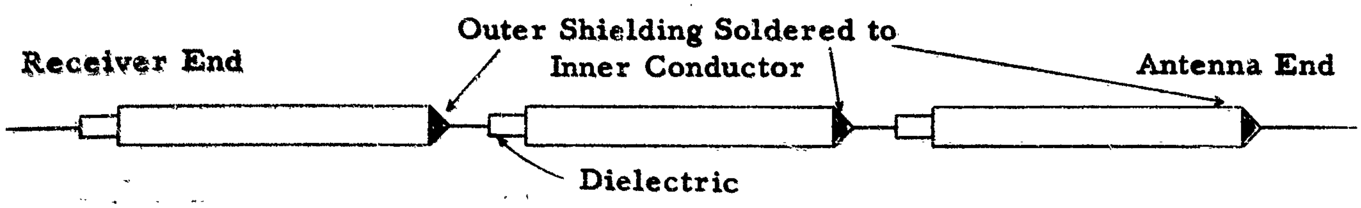


Fig. 3.1.4.2-E Three Section Parallel-Resonant Series Wave Trap

3.2 DESIGN CONSIDERATIONS APPLIED TO AIRCRAFT COMPONENTS FOR MINIMUM GENERATION OF INTERFERENCE

Radio interference originates from the operation of the components of aircraft systems. Consideration of the aircraft system itself is essential only to establish techniques to prevent the interference which is generated by the components from

being transmitted by radiation, conduction, or coupling to various other susceptible receivers in the aircraft. These system considerations are treated in detail under Paragraph 3.3. The ideal method of eliminating the effects of unwanted signals is to design all components in such a way that no unwanted signals are generated.

Source suppression is by far the best method of controlling interference in most cases and should be applied whenever possible. Nevertheless, in some components the generation of signals is inherent to their normal function and source suppression cannot be employed. This is true for all transmitters where the signals appearing on the transmitter antennas are the desired result. Here the interference problem is primarily a system design consideration involving mounting and location of the antenna, and taking advantage of the shielding effects afforded by the aircraft structural members and metallic skin. However, source suppression techniques can be employed in the transmitter case design and in the elimination of harmonics appearing on the antenna to reduce the transmitter interference problem considerably by eliminating various unwanted signals associated with the generation of the desired output signal.

Source suppression is desirable from several standpoints other than that of interference-free design. Aircraft maintenance is one important reason for utilizing source suppression wherever practicable. The resulting decrease in required shielding greatly reduces the electrical maintenance problem. This is especially true when the aircraft is to be operated under combat conditions. Flak or gunfire damage at any point along a conduit would require replacement of an entire rigid conduit assembly and would considerably prolong the non-operational status of the aircraft while under repair. When flexible conduit is utilized for shielding purposes, as is the case in all late model aircraft, the maintenance problem is somewhat less severe. Nevertheless, damage to any required shielding increases the time and cost of repairs.

Since source suppression can best be employed in the original component design, the techniques described in the following paragraphs are of major importance. Design engineers should be thoroughly acquainted with this material to insure good interference-free components. It is of utmost importance that all electrical components, regardless of their function or location within the aircraft, be treated as potential sources of interference.

3.2.1 MOTIONAL SOURCES

Most of the interference generated by the operation of electrical devices in an aircraft originates in commutator-type machines and arcing contacts. These include all the motors, generators, vibrators, relays, and switches which perform functions essential to the operation and control of the aircraft. Modern aircraft have become a maze of such motional sources of interference. Figure 3.2.1 shows the location in a typical very heavy bomber of various electrical components, and the functions performed by these components. This clearly indicates that these components located in all parts of the airframe are essential to the operation of the aircraft and the associated interference problem cannot be avoided.

When Figure 3.3-C is compared with Figure 3.2.1 it becomes quite apparent that every electrical component in the aircraft may possibly be mounted sufficiently close to susceptible components or component wiring, to introduce disturbing signals

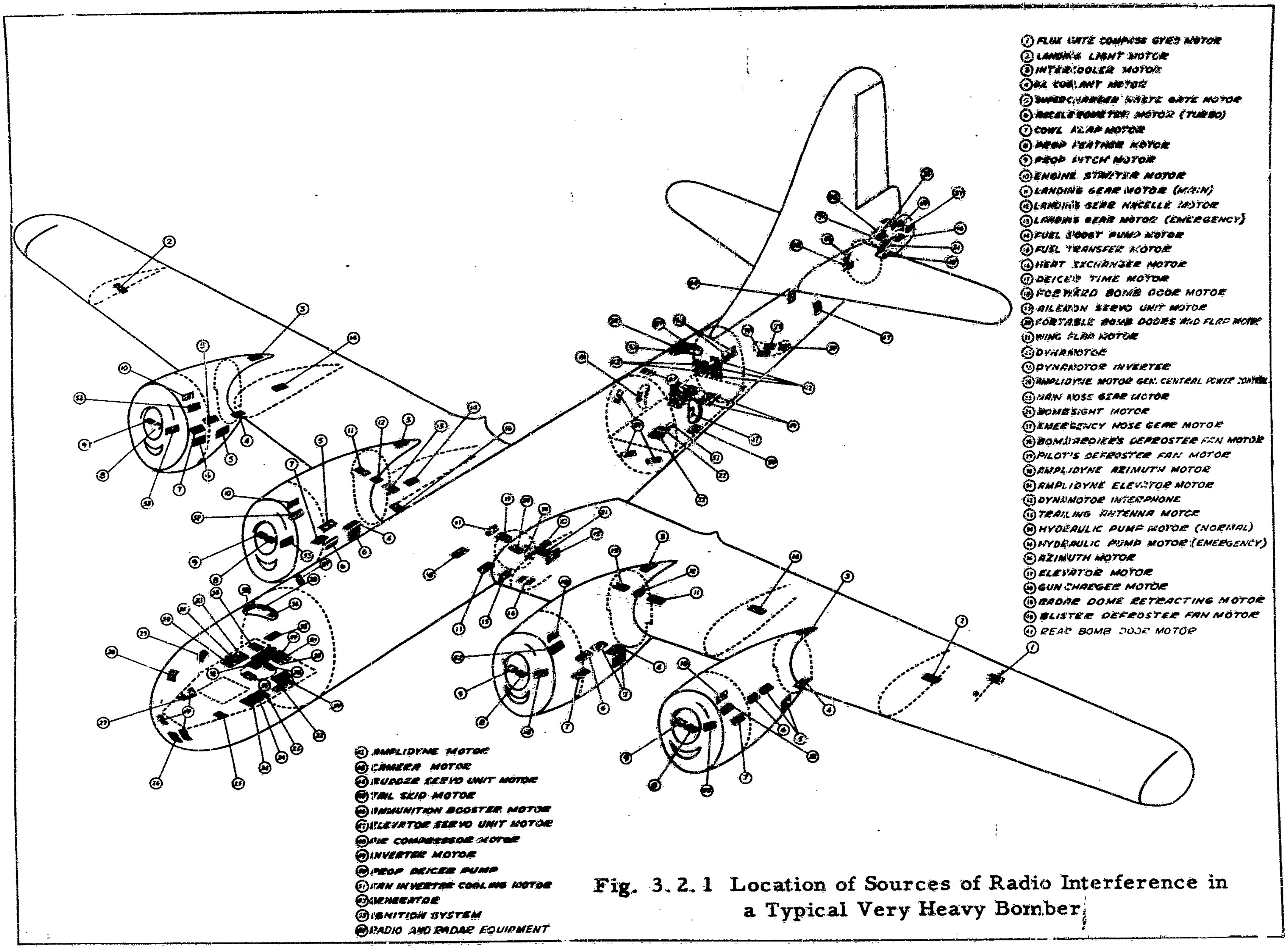


Fig. 3.2.1 Location of Sources of Radio Interference in a Typical Very Heavy Bomber

into these units and adversely affect their operation. Consequently each and every motional source must be designed so that inherent electrical transients are confined to the unit itself and not permitted to enter other components either directly or indirectly and cause radio interference.

3.2.1.1 ROTATING MACHINERY

Of all the sources of interference commonly encountered in aircraft, rotating electrical machinery constitutes the largest single group. It is, therefore, necessary to be particularly careful in the design of such machines if interference-free operation is to be achieved.

The first point to be stressed is that a good, clean design is usually also the design least likely to cause interference. What needs to be remembered is that many considerations that seem secondary when interference is neglected become primary in the light of interference problems. Such obvious points as symmetry in the windings, mechanical and electrical balance, accuracy of machined parts, and tolerances in the assembly acquire an entirely new importance when interference problems are considered from the beginning in the basic design.

In the following paragraphs, certain considerations that are common to many types of rotating machines will be discussed first. Then the detailed design of special types of machines will be taken up in turn.

3.2.1.1.1 BRUSHES

In all types of electrical machines (except certain types of induction motors) electrical contact must be made between two conducting surfaces that are in relative motion. Such contact is normally made by brushes sliding on a metallic surface. As explained in Paragraph 1.3.2.1, this is always accompanied by the generation of interference. In DC machines having good commutation, most of the interference is directly attributable to the sliding brush contact. This so-called brush interference may be reduced by careful consideration of the following factors in the design or choice of brushes and of the metal surface in contact with the brush face.

(a) Brush Pressure. Interference generated decreases with increasing brush pressure at all frequencies as shown in Figure 3.2.1.1.1-A. As the brush pressure is increased, the unit pressure over the entire brush face in contact with the metal surface is maintained more constant and uniform, thereby reducing the variation in contact resistance across the sliding surfaces which produces what is known as surface contact transients. In addition, the possibility of brush bounce which causes severe arcing and transients is reduced. Since the amount of brush vibration and chatter increases with the peripheral speed of the sliding metal surface in contact with the brush, the brush pressure selected for any particular application should at least be adequate to damp-out the vibration expected at the designed peripheral speed. Increased brush pressure increases the rate of wear, but the necessity of more frequent replacement should be considered a reasonable compromise for the sake of decreased interference.

(b) Current Density. Interference generated increases with increased current

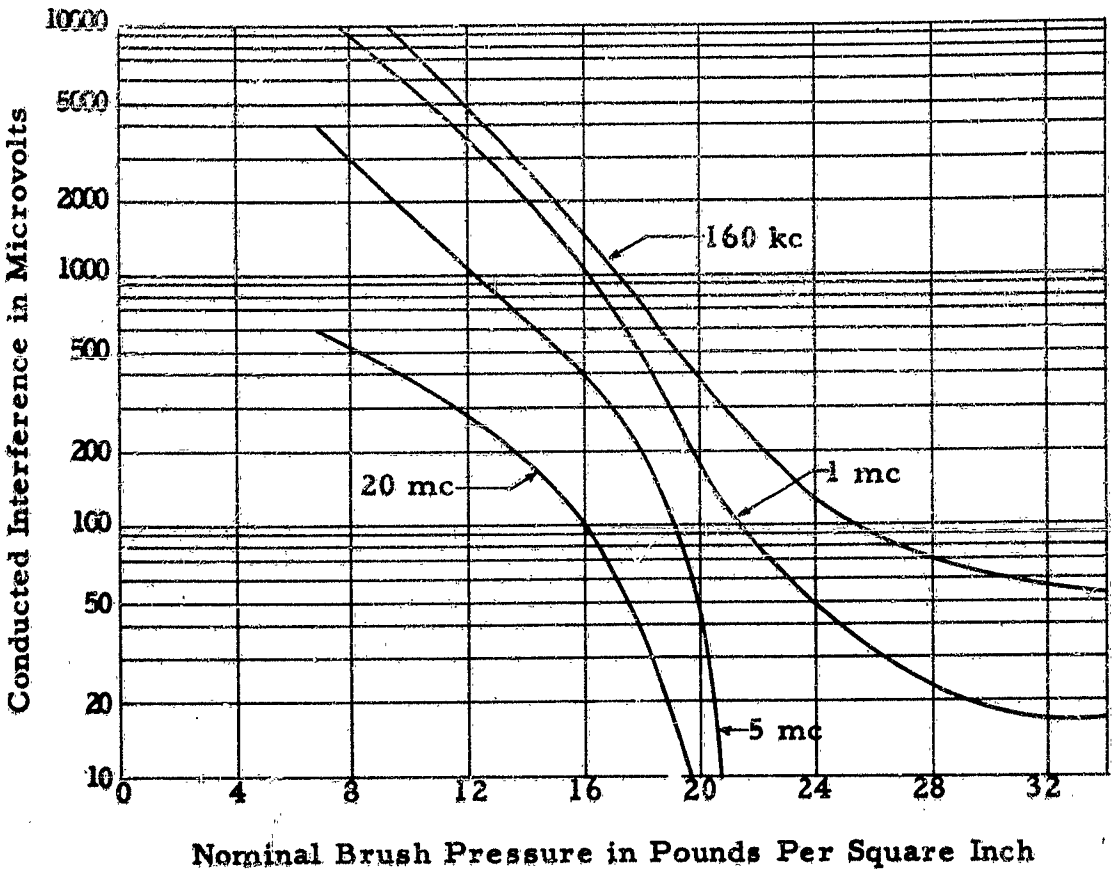


Fig. 3.2.1.1.1-A Effect of Brush Pressure on Generation of Interference at Various Frequencies

density as shown in Figure 3.2.1.1.1-B. As the current density is increased, more heat is generated in the contact resistance and the temperature of the brush surface in contact with the commutator increases. This increase in temperature hastens the formation of an oxide film of considerable thickness on the sliding metal surface. Rapid variations in sliding contact resistance due to irregularities in the oxide film modulate the direct current and may give rise to radio interference. The magnitude of interference so produced is increased as the current density, hence voltage drop, across the brush face increases. Therefore, somewhat larger brush surface area should be provided than is demanded only by consideration of the dissipation of heat and losses due to mechanical friction. However, if too low a current density is used, non-uniform grooves or threading develop on the metal surface or commutator, and frequently a high friction coefficient occurs which sets the brushes into a noisy chatter. General design practice calls for a contact current density of 55 - 65 amperes per square inch at full-load rating for electrographitic carbon brushes, and 65 - 90 amperes per square inch for metal-graphite brushes.

(c) Surface Treatment. Interference generated is materially decreased by treating the brush surface with a colloidal graphite material. This treatment also decreases the mechanical friction and the voltage drop across the brushes. The metal surface upon which the brushes bear deserves special consideration because of the prolonged effects of wear and high

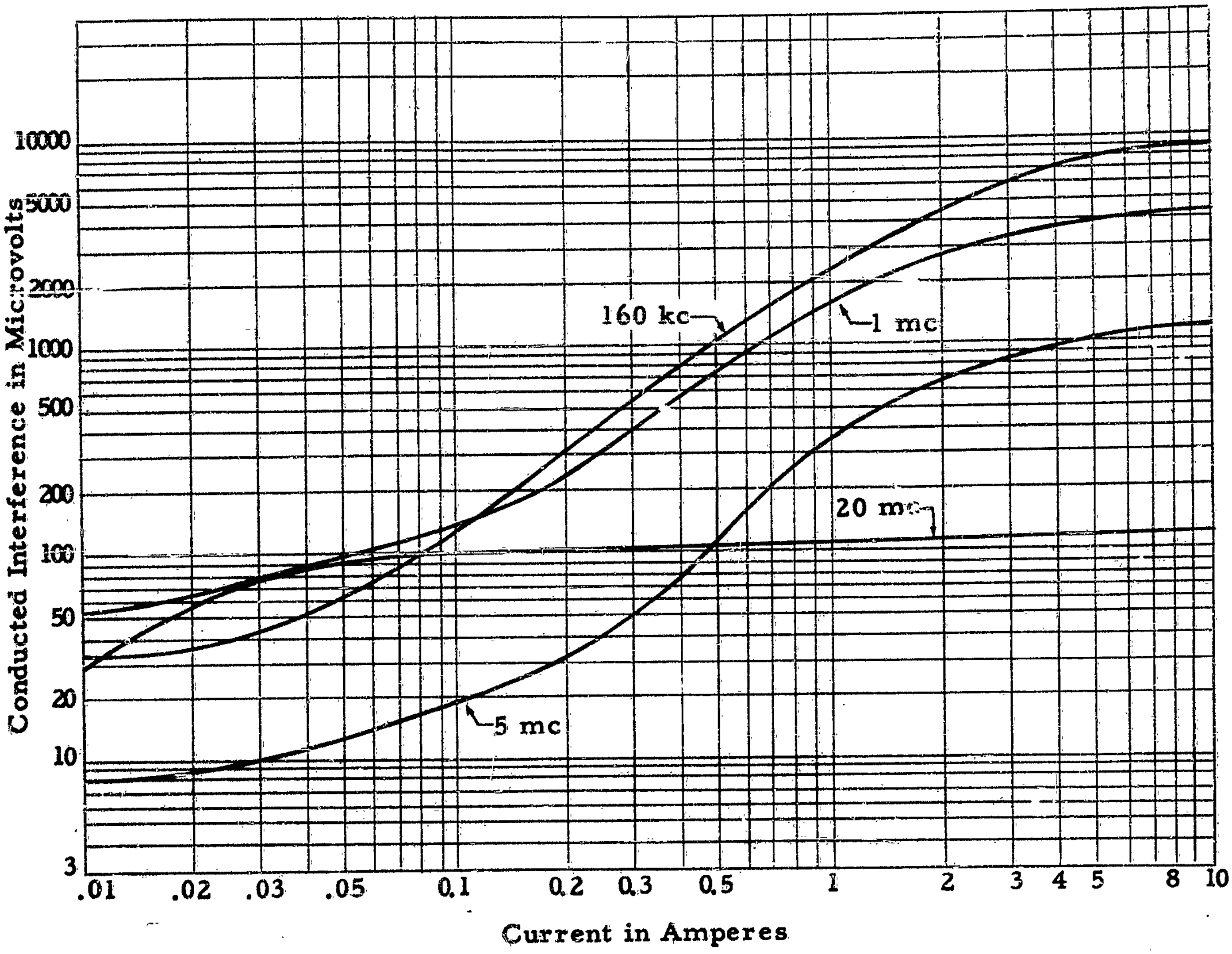


Fig. 3.2.1.1.1-B Effect of Brush Current on Generated Interference at Various Frequencies

temperatures to which it is subjected. A newly finished metal surface develops an oxide film within several hours. For instance, in the case of a copper commutator in contact with a carbon or graphite brush, a layer of copper oxide, mixed with carbon particles from brush wear, forms on the commutator. The presence of this copper oxide film introduces unidirectional electrical properties (polarity effects) consistent with the well-known copper-oxide rectification. The oxide layer displays a non-linear resistance of higher value to a brush used as a cathode than to one used as an anode. The cathode brush passes current in discontinuous high current density surges. Approximately ten times as much radio interference may come from the cathode brush as from the anode brush. Promising experimental results have been obtained by plating the copper commutator with chromium to a thickness of about one mil, reducing the interference observed from a cathode brush to that of the relatively quiet anode. The somewhat higher resistivity of the chromium plating does not cause appreciable loss as regards heat dissipation at the contact inasmuch as the commutator-film power loss is relatively large. The thinness of the chromium oxide

layer and the fine division of the finished chromium surface seem to account for its excellent performance. This performance is maintained because the hard chromium surface avoids threading and grooving of the commutator. Wear rate and sliding friction of many brush materials on chromium is of the same order as that for copper.

(d) Brush Resistivity. Interference generated is less for brush materials of lower resistivity. General design practice for good performance is to use an electrographitic carbon brush with 0.0015 to 0.0025 ohm specific resistance in machines operating above 50 volts, and a metal-graphite mixture for brushes in machines being used at less than 50 volts. Silver, copper, or cadmium impregnated graphite is available in the form of low-resistance metal-graphite brushes. The final selection of a brush material is actually a compromise after consideration of all the mechanical and electrical properties which the brush must possess. Nevertheless, the material of lowest resistivity which still satisfies the other requirements for good functional performance should be the preferred choice. The resistivity of brushes used for commutation should also be in accord with the requirements for good commutation as noted in Paragraph 3.3.1.1.2. On the other hand, considerable leeway is permitted in the choice of design and brush material for slip-ring applications since no switching action is involved. The use of low resistance brushes can be applied to good advantage. For instance, the use of a low-resistance brush composed of many flexible metal contacts sliding on a chromium surfaced slip-ring gives radio interference reduction substantially greater than 10 to 1 over normal brush contacts.

(e) Altitude Treatment. Brush and slip-ring or commutator devices used in aircraft rotating electrical devices must be designed for high altitude operation. Brush wear at high altitudes is much more severe than at ground level. This has been attributed to the rarified atmosphere and lack of moisture at high altitudes. Under such conditions a layer of oxide does not readily form on the sliding metal surface. While the presence of this film is undesirable from a radio interference standpoint because of the resulting variations in contact resistance, its presence does reduce wear by serving as a lubricant. Severe wear, especially if it is not uniform, is just as bad from a radio interference standpoint as the presence of a film layer. Consequently, special brushes are used which contain the ingredients necessary for a film layer to form. Impregnated brushes have been developed for this purpose with built-in lubrication and/or oxygen supply. The most successful of such impregnated brushes are those utilizing barium compounds in varying percentages. It should be noted that hermetical sealing of units produces conditions within the container which approximate those at high altitudes. Hence special brushes should be used in such units also.

3.2.1.1.2 COMMUTATION

Commutation, as explained in Paragraph 1.3.2.2, is essentially a switching action, and as such is normally accompanied by interference producing transients - called break transients. This commutation interference is apart from the brush

interference, or surface contact transients, explained in Paragraph 3.2.1.1.1. However, commutation is always achieved b brushes bearing on a commutator so that both contribute to the interference genera. d by commutator devices. This combined interference is commonly called "motor hash" - a poor term to use in design practice since no distinction is made between brush and commutator interference. For design purposes, such a distinction should be made since the techniques applicable to suppressing each are different. The reduction of brush interference necessitates providing a low, non-varying contact resistance between brush and sliding metal, while reducing commutation interference necessitates the use of special features designed to provide as smooth a transition as possible from one value of current to another. Consideration of the requirements for good commutation may in some cases conflict with the choice of technique used to suppress brush interference. In general, good commutation deserves first consideration.

For this reason, machines requiring commutation are doubly troublesome from a design standpoint. The design engineer should first determine whether or not a machine requiring commutation is absolutely necessary for a particular application. If, for example, an induction motor can be substituted for a direct current motor, it should be done even at some sacrifice of cost, ease of wiring, or ease of control. Unfortunately, such a substitution is not always possible, mainly because of the high starting torque of certain direct current motors, which cannot be duplicated by alternating current machines except by the addition of bulky and complicated devices. Therefore, where commutation is judged to be essential, special attention must be given to incorporating design features which will minimize the interference generated.

Commutation interference itself, aside from any consideration of brush interference, may be reduced by the use of (1) interpoles, (2) compensating windings, (3) laminated brushes, and (4) careful machining techniques to insure clean, symmetrical mechanical design. All these techniques are intended to smooth out the commutation break transient as much as possible. Figure 3.2.1.1.2-A shows a typical oscillographic voltage and current trace of an armature coil undergoing commutation in a machine which employs none of these special design techniques. The diagram actually illustrates an example of poor commutation where strong interference is generated.

At point 2 in this figure, the current reversal in the coil nominally begins with initial contact of the leading edge of the brush with the commutator bar. At point 4, the reversal of current (I) from plus to minus is complete although the trailing edge of the brush broke contact with the bar at point 3 where a break transient is visible on the voltage wave (E). Therefore the commutation break transients are seen to originate at the trailing edges of the brushes, i.e., at the end of the commutation period for each armature coil. The rise in current evident in Figure 3.2.1.1.2-A between points 1 and 2 on the current wave (I) is the result of a break transient produced in the coil immediately adjacent to and in advance of the coil undergoing commutation in the figure. It is evident that transition in current from point (1) to (4) is far removed from the ideal commutation noted in Paragraph 1.3.2.2 and the generation of severe interference is to be expected. In order to reduce the generation of commutation interference the steepness of the break transient should be reduced and the initial rate of change of current in the armature coil should be greater. The following techniques are employed to achieve this goal.

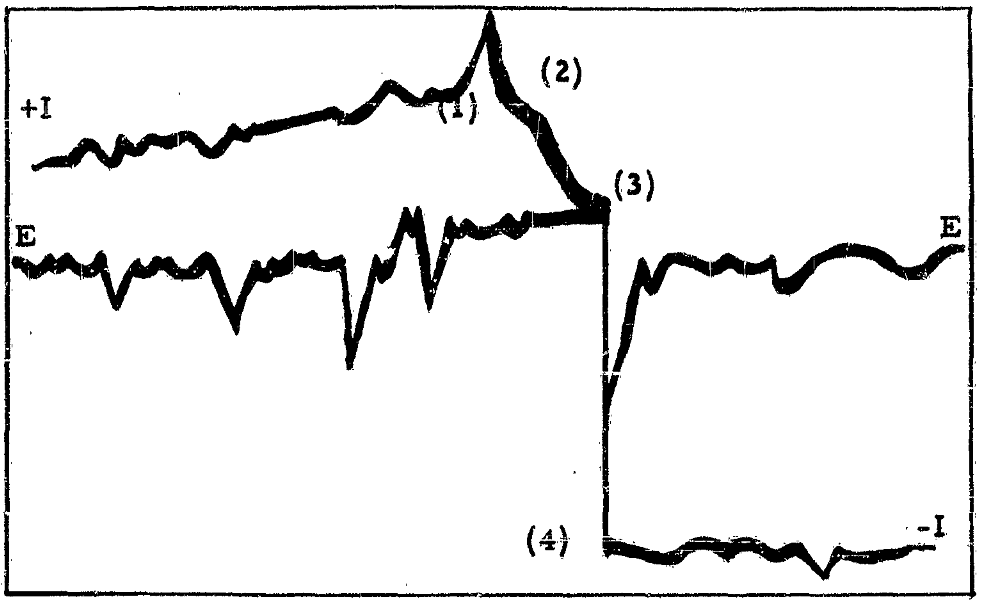


Fig. 3.2.1.1.2-A Voltage and Current Oscillogram During Commutation Period for an Armature Coil

1. Interpoles. The best way of improving commutation is by the addition of interpoles. The main function of interpoles in DC machines is to (1) counterbalance the self-induction of the armature coils during the commutation period, and (2) reduce the induced voltage in the armature coils resulting from the coils cutting fringing flux from the pole pieces during the commutation period. Hence, the use of properly designed interpoles would produce a more rapid change in the armature coil current at the beginning of the commutation period and thereby reduce the steepness of the break transient at the end of the commutation period. Interpoles are usually not practical on small machines because of the added weight and lack of space. Yet careful attention should be given to the possibility of using interpoles even though contrary to normal practice.

2. Compensating Windings. When interpoles cannot be used, compensating windings in the pole pieces will produce the same effect as interpoles. Compensating windings are rarely used because of the expense involved in cutting the slots into the pole pieces and inserting the windings. In many cases, it will be found that the additional expense is justified in view of the decrease in the generation of interference.

3. Laminated Brushes. If practical limits prevent the use of either interpoles or compensating windings, aircraft-type machines usually rely on so-called resistance commutation. This scheme should reduce the current in the receding bar to zero at the time the bar leaves the trailing edge of the brush. Actually, as Figure 3.2.1.1.2-A indicates, the current does not reach zero at the appropriate time and a steep break transient is produced when the bar leaves the brush.

Instantaneous values of coil current during commutation should depend only upon the ratio of the contact drops of the brush to the commutator bars connected to each end of the armature coil being commutated. Since the self-inductance of the

armature coil tends to oppose the change in coil current, the armature current does not divide in the same ratio as the contact drops of the brush to the commutator bars. Circulating currents flow in the coil undergoing commutation by way of the commutator bars and brush, and the rate of change of coil current is not constant, being very slow initially in the commutation period as shown near point 2 in Figure 3.2.1.1.2-A. If the brush resistivity is large enough to reduce circulating currents in the beginning of the commutation period, the initial current reversal would be at a faster rate. Hence, low brush resistivity, while desirable from the standpoint of reducing brush interference as mentioned in Paragraph 3.2.1.1.1, is undesirable from the standpoint of good commutation.

Lowering the resistance presented to the approaching commutator bar and raising the resistance presented to the receding bar would favor the flow of load current in the approaching bar and also would introduce higher resistance to the circulating current. The use of laminated brushes with brush materials of different resistivity produces such control and hence gives a more linear current reversal. Promising experimental results have been achieved with a brush consisting of segments of different resistivities, the trailing segment having the highest and the leading segment having the lowest. The segments are well insulated from each other.

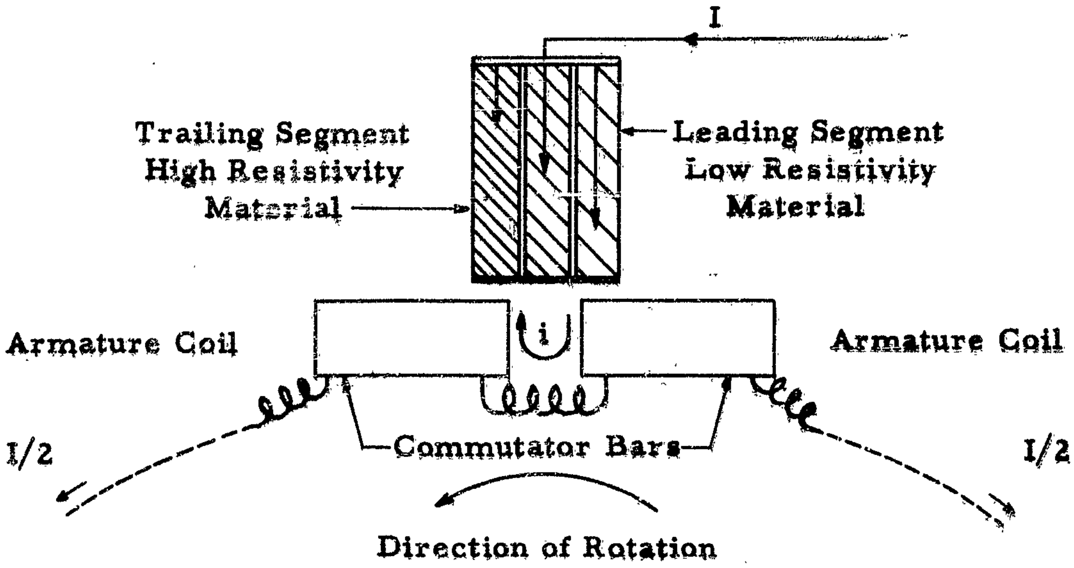


Fig. 3.2.1.1.2-B Commutation of an Armature Coil by Using Laminated Brushes

The ideal operation of laminated brushes is indicated in Figure 3.2.1.1.2-B. In this diagram, a three-lamination brush is bridging the commutator bars attached to each end of the coil undergoing commutation. The brush resistance increases as the commutator bar progresses from the leading edge to the trailing edge of the laminated brush. The segments are insulated from one another by some suitable glue and electrically connected only at the end in contact with the brush spring. Hence, circulating currents resulting from the self-inductance of the coil under commutation and from the coil cutting fringe flux from the pole pieces must flow through the entire length of two brush laminations whose total resistance is much greater than that presented by a direct path across the face of the brush as would occur with a non-laminated brush. Consequently, circulating currents are reduced early in the commutation period and a more desirable division of current through the two adjacent

commutator bars is achieved. A more linear coil current reversal is produced and break transients are considerably reduced.

Another consequential advantage to the use of laminated brushes is that good commutation can be achieved over a fairly wide range of brush positions relative to the magnetic neutral, so that this position becomes less critical and less dependent on the armature current.

The Details of Laminated Brush Designs are discussed in Appendix XIV. The lamination design may be summarized as follows.

(a) Two, or at most three, laminations give best performance.

(b) The thickness of the leading edge lamination of a two lamination brush should be about 90 percent of the total brush thickness and its resistivity should be as high as allowable by consideration of heat dissipation.

(c) For application to small machines, good performance is obtained when the resistivity of the trailing edge lamination is about 15 times that of the leading edge.

(d) The trailing lamination should be thick enough to avoid mechanical weakness.

(e) The glue thickness should be sufficient to provide electrical insulation and to avoid the formation of a smear of conducting particles from brush wear on the rubbing edge which would short-circuit the laminations. (A thermosetting glue of six-mil thickness has been found satisfactory.)

Reference to Figure 3.2.1.1.2-A shows a series of radio interference transients before point 1 where commutation begins. These are the surface contact transients mentioned in Paragraph 3.2.1.1.1. For the case illustrated in this figure the break transient at point 3 is relatively large compared to the surface contact transients. This is usually the situation observed for poor commutation and is accompanied by trailing edge burning of the bars or brushes. Radio interference reduction by the use of laminated brushes is generally limited to such cases where the break transients are relatively large. In other cases, interference reduction can be better achieved by employing one or more of the techniques noted for the reduction of brush interference.

4. <u>Machining Techniques</u>. Any other design feature that improves commutation in general will also reduce the generation of interference. Considerations such as uniform distribution of armature coils with respect to the commutator bars should be given careful attention in the light of the requirements for good commutation. Careful machining techniques are necessary to insure good electrical and mechanical symmetry. For instance, commutator interference can be reduced by grinding the commutator precisely about its true rotating axis. In a typical case of a very troublesome DC generator, (a) the armature was chucked on the shaft centers, and the commutator was ground. The interference reduction resulting from this operation was in the ratio of approximately 2:1; (b) the same commutator was then ground, with the armature chucked in the generator bearings. This resulted in an interference

reduction by a factor of about 25:1. Operation (a) represented the expected rotation axis, while operation (b) represented the true rotation axis.

3.2.1.1.3 USE OF CAPACITORS AND FILTERS

Even with the best design, some interference will still be generated at the brushes and during commutation. To prevent this from being conducted to other equipment, capacitors or filters should be incorporated in the original design. In some cases, a simple capacitor of from 0.05 to 1 microfarad, depending on the size of the machine and the amount of interference generated, connected directly across the brushes, is sufficient. In other cases, a complete filter in the output leads may be required.

In present day practice, machines are often constructed without filters, and a filter is added later if excessive interference is produced. These filters are often found ineffective because, once the machine is completed, the brush terminals are not accessible, so that leads of considerable length may be required to connect the filter, or the filter is grounded to the frame of the machine which serves as ground for one of the terminals. As was pointed out in Paragraph 1.8.2.3, extremely short and direct connections are very important for the effectiveness of a filter. The great advantage of incorporating filters or capacitors into the original design is that connections can be made at the point where the suppressing action is most effective. Much is to be gained, for example, as far as avoiding radiation and capacitive coupling is concerned if the interfering currents can be kept entirely out of the frame of the machine. In addition, a capacitor connected directly across the brushes will be much more effective than one connected from the output lead to the frame, but located several inches away from the brushes.

In the installation of filters or capacitors, the consideration of good bonding (see Paragraph 3.1.3) is extremely important. In many actual cases, recent experiences have shown that a filter or capacitor was ineffective only because some protective coating was not removed so that no electrical contact was made between the filter and its base. Where installation directly at the brushes is impossible, great care should be exercised in shielding the input leads from the brushes to the filter to the point of entry. Obviously, care should be given to the prevention of any sort of coupling between the "clean" output leads and the "noisy" input leads to a filter.

Filters for 400 cycle alternating current machines require special attention. An ordinary capacitor may draw too much current at the fundamental frequency to be practical. In that case, it is necessary to design a low-pass filter at the proper impedance level with a cut-off frequency of about 600 cycles per second according to the design equations of Appendix VII.

3.2.1.1.4 SHIELDING

The brushes and brush leads are the most likely regions from which interference may be radiated or coupled with other circuits. Therefore, unless the entire machine is completely enclosed in such a way as to be adequately shielded, the brushes, brush holders, and brush leads should be shielded as completely as is possible without disturbing their normal functioning. Complete shielding of the entire

machine should be used, when practical, in the case of small direct-current machines such as dynamotors, which generate a considerable amount of interference even with the best design. However, the shaft provides a path for interference since it must penetrate the shielding. The shielding of the shaft is done in any of the following ways: using a nonconducting fiber coupling in the shaft, or using a conductive packing for the bearings where the shaft penetrates the shield, or using an additional set of brushes for the shaft near the place where it passes through the shielding.

3.2.1.1.5 SERIES DIRECT-CURRENT MOTORS

Series direct-current motors are most frequently used in aircraft because of their high starting torque. Many of them have split field windings so that they can be reversed quickly without changing more than one connection.

The considerations of Paragraphs 3.2.1.1.1 and 3.2.1.1.2 are applicable to series direct-current motors. They may require filters though first consideration should be given to a good, clean design, which makes filters unnecessary. If filters are used, the series winding may be utilized as their series element so that only one or two additional capacitors are required. The capacitor should be placed as shown in Figure 3.2.1.1.5-A because first only one instead of two capacitors are required in case of a split field winding, and, secondly, this position leads to greater attenuation as proved in Appendix IX. If this is not practical because of lack of space, two separate capacitors must be used as shown in Figure 3.2.1.1.5-B. In several experimental cases, the attenuation was increased by splitting the series field and adding the capacitors as shown in Figure 3.2.1.1.5-C. In extreme cases, a pi-section should be constructed by using capacitors simultaneously in all three places. Consideration should be given to isolation of field leads when utilized as a part of a pi-section filter. These leads should be shielded between the armature and field connection. The power input lead should be routed to be as remote as possible from all other leads. The size of the capacitors is determined by the size of the machine and the amount of interference generated, as explained in Appendix VII, and will normally be about 0.05 to 0.5 microfarads.

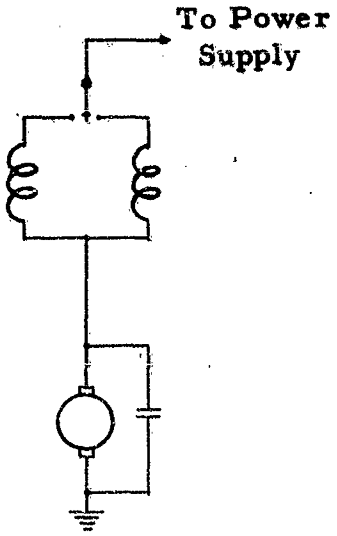


Fig. 3.2.1.1.5-A Location of Capacitor in Series Motor

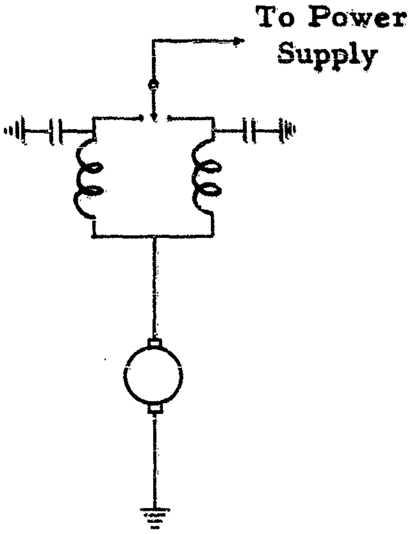


Fig. 3.2.1.1.5-B Alternate Location of Capacitors in Series Motor

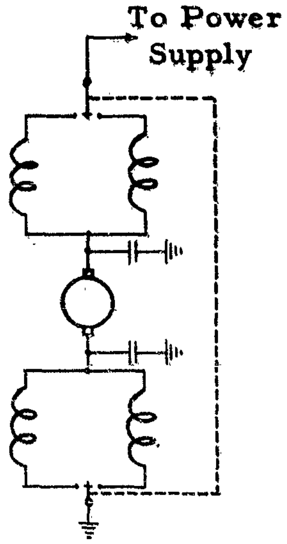


Fig. 3.2.1.1.5-C Location of Capacitors in Split Series Field Motor

3.2.1.1.6 OTHER DIRECT-CURRENT MOTORS

The same considerations also apply to all other direct-current motors. In a compound motor, the series field may be used as part of the filter as in the series motor as shown in Figure 3.2.1.1.6-A. But in a shunt motor, a complete filter must be installed as shown in Figure 3.2.1.1.6-B because the shunt field cannot be utilized for that purpose. In most applications, dynamotors are small units placed directly at the location where their output is used. Since they have two commutators, they are fairly effective interference generators, but filters, other than comparatively small capacitors across the brushes, are not practical because of their weight and size. The most practical solution for these small units is complete shielding in accordance with the considerations of Paragraph 3.1.2.

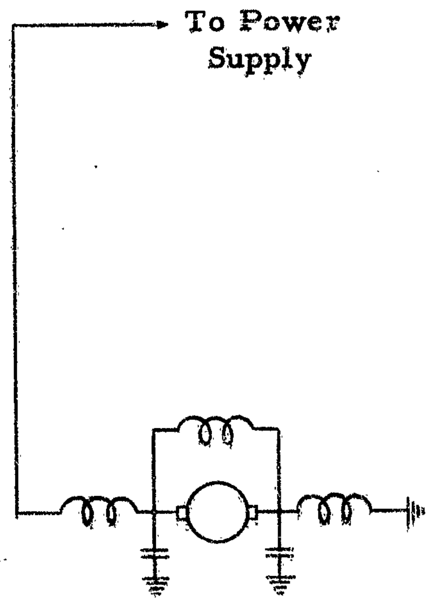


Fig. 3.2.1.1.6-A Installation of Capacitors in Direct Current Split-Field Compound Motor

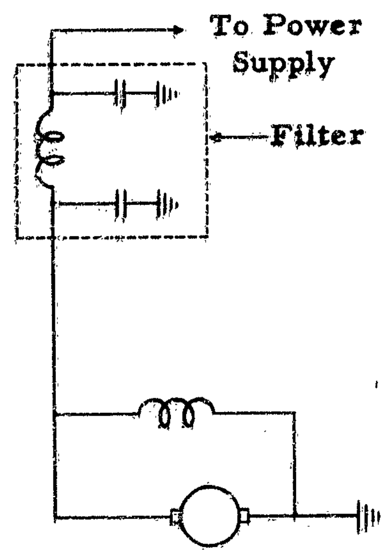


Fig. 3.2.1.1.6-B Installation of Filter in Direct Current Shunt Motor

3.2.1.1.7 DIRECT-CURRENT GENERATORS

Direct-current generators are basically very similar to direct-current motors and the same design considerations apply. Since the field is normally fed from the output of the generator itself, this would be classified as a shunt machine, and the field cannot be used as part of the filter. This is a serious restriction although in most aircraft there are only one or at most a few generators, and these are large ~~so the additional size of a fairly heavy filter can be prohibitive.~~ Adequate filters should be incorporated in the positive output lead as close to the brushes as practical in those cases where the airplane structure is used as negative return path. Even here, direct connection of the filter ground terminal to the negative brush is preferable to grounding it to the frame only. Careful attention must be given to proper bonding as pointed out in Paragraph 3.2.1.1.3. If the negative lead is not grounded, a filter must be used in each output lead and a capacitor should be connected directly across the brushes. It should be remembered, of course, that filters are used only as a last resort, and that a good design may make all filters unnecessary.

3.2.1.1.8 ALTERNATORS

The considerations of Paragraphs 3.2.1.1.1 through 3.2.1.1.4 apply to all alternating current generators. In addition, careful attention must be given to the prevention of the generation of harmonics. Production of as pure a sine wave as possible is one of the important "normal" considerations in the design of alternators. But this requirement acquires a new importance and is put to a much more severe test when the generation of radio interference is considered. A comparatively minute harmonic content might be quite tolerable from all points of view except that of radio interference.

Alternators usually use direct-current exciters to provide the necessary magnetic field. These should be designed in accordance with the recommendations of Paragraph 3.2.1.1.7 except that their size usually does not warrant the use of a complete filter, and a single capacitor connected across the brushes will normally provide sufficient filtering action.

To reduce the generation of harmonics, special attention should be given to the following items:

(a) Flux distribution. The most important factor determining the wave form of the generated voltage is the distribution of the magnetic flux around the periphery of the armature. Sinusoidal distribution may be achieved by chamfering the pole tips or skewing the pole faces.

(b) Symmetry. For a perfectly symmetrical machine, all even harmonics automatically disappear. Therefore, special care must be exercised in constructing identical pole pieces, making the yoke and armature perfectly symmetrical, producing a perfectly uniform winding on the armature, and avoiding all other irregularities.

(c) External Connections. In a three-phase alternator, the third harmonic and its multiples disappear at the terminals except when the machine is star connected and has its neutral grounded, in which case third harmonics are present in the voltage from any phase to neutral. Hence, this connection should be avoided, or else special attention must be given to the prevention of the third harmonic and its multiples.

(d) Chord factor. The harmonics generated may be considerably reduced by the choice of a suitable chord factor. If the difference between the pole pitch and the coil pitch is θ electrical degrees, the chord factor for the n^{th} harmonic is $\cos(n\theta/2)$: A value of 30° is often recommended for θ because this greatly reduces the chord factors for the 5th and 7th harmonics while affecting the fundamental very little.

(e) Distribution factor. If the winding is distributed over several slots per pole per phase and there are m slots per pole per phase, the distribution factor for the n^{th} harmonic is $\frac{\sin(mn\theta/2)}{m\sin(n\theta/2)}$, where θ is the slot pitch in electrical degrees. This decreases with increasing m and should be chosen so as to eliminate the lowest harmonic not eliminated by any of the devices mentioned in (b), (c), and (d).

(f) Tooth ripples. The generation of tooth ripples can be greatly decreased by skewing through one slot pitch either the pole shoes or the armature slots. Also, tooth ripples may be eliminated altogether by making the number of armature slots per pole-pair an odd number. For, in this case, the chord factors for the harmonics that are contained in the tooth ripples reduce to zero.

Incorporating the above considerations into the original design of an alternator produces a machine which generates a minimum of interference. Such practice is highly desirable since less demands have to be made on filtering techniques to bring the remaining conducted interference within tolerable limits.

Complete filters are not usually required in the output of well-designed alternators. In a three-phase machine, capacitors should be connected directly across each set of brushes as shown in Figure 3.2.1.1.8, but their value must be somewhat smaller than that recommended for direct-current machines because the current at the fundamental power freuqency must be kept low. If filters should be required, they must be designed as low-pass filters with a cut-off frequency of about 3/2 times the fundamental power frequency, and one is required in each lead except the ground lead.

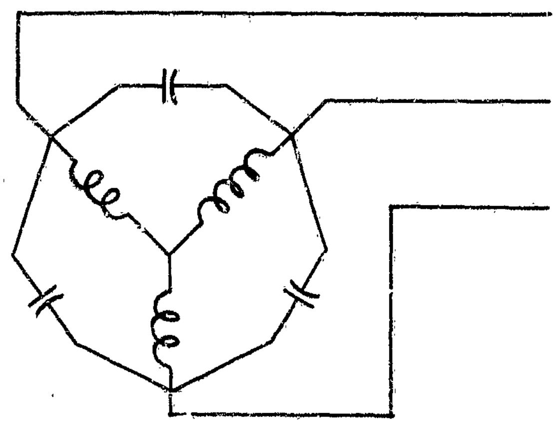


Fig. 3.2.1.1.8 Installation of Condensers In Three Phase Alternator With Ungrounded Neutral

Even a well-designed machine will radiate some interference which must be prevented from leaking out of the alternator housing. This necessitates a casing which acts as a perfect shield and is designed to permit a good bond to the airframe when installed in an aircraft. The additional requirement of adaquate ventilation, especially for a machine in continuous operation, makes it difficult to design a perfect shield. This difficulty can be overcome by providing ventilation through tubular air vents which act as wave-guide attenuators to the interference.

Consideration of the problems encountered in a typical motor-alternator unit will illustrate some design procedures that are useful in bringing the generated interference within tolerable limits. The example chosen and discussed below is actually a case of a motor-alternator that was well-designed from a functional standpoint and contained a filter as an integral part of the unit. However, from the radio

interference standpoint, the original design did not adequately provide for good shielding and bonding. The modifications made to the unit could easily have been incorporated into the original design.

The motor-alternator was subjected to extensive tests which indicated that, after obtaining satisfactory reduction of conducted interference by the use of a filter in the DC power line, excessive radiated interference was present in the medium high and the ultra high frequency ranges. A unique technique for the attenuation of radiated interference in motor-alternators was developed by the Navy. A circular section containing the ventilating louvres was cut from the commutator end-bell of the housing and replaced by a similar section having tubes of small diameter, which act as wave-guide attenuators as explained in Paragraph 3. 1. 2. 2, and at the same time provide the necessary ventilation. This section of the housing before and after modification is illustrated in Figure 3. 2. 1. 1. 8-A. The inside diameter of each tube

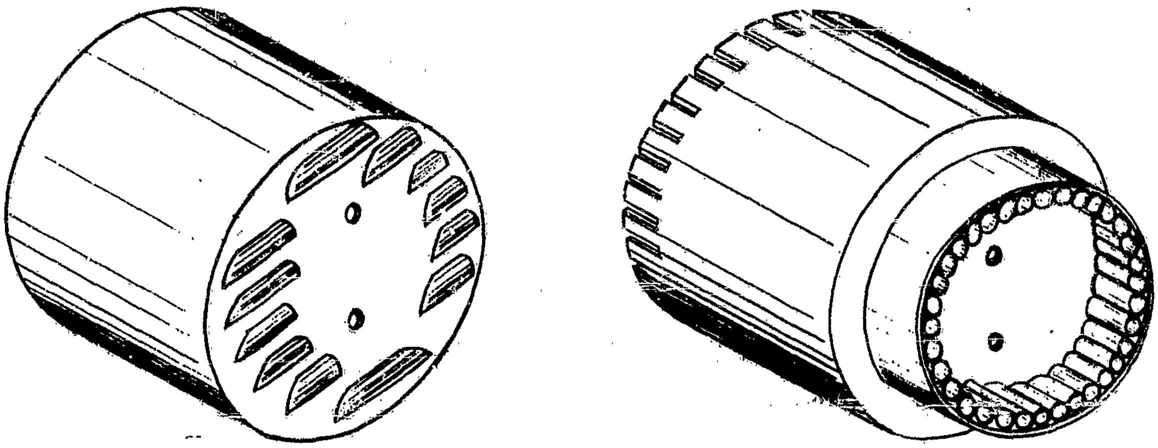


Fig. 3. 2. 1. 1. 8-A Section of Motor-Alternator Shield Before and After Modification

used was 19/64 inch, and its length was in excess of 57/64 inch. The number of tubes used was determined by the requirement that the total cross-sectional area of the openings should not be less than 1.6 square inches to assure sufficient ventilation. Copper was used for this modification but any metal having high conductivity, such as aluminum or brass, could be used provided that all joints are capable of being soldered or welded so that no opening, other than that of the tubes, exists. Paragraph 3. 1. 2. 2 emphasizes the necessity for avoiding openings in the shield. In addition, a metal collar no less than 1/32 inch in thickness was placed around these tubes to provide rigidity and protection. The rim of the commutator end-bell was slotted, and the resulting "fingers" were cleaned and polished as well as sprung slightly inward to assure continuous and positive metal-to-metal contact of the friction fitting that joins the commutator end-bell section to the rest of the alternator housing. Such modification to the casing of the alternator provides for adequate attenuation of the radiated interference. In general, the use of standard ventilating louvres should be avoided since the leakage of high frequency interference through such openings can be severe.

Since the example under consideration is a motor-alternator unit, the motor itself must be designed for interference-free operation. Attenuation of the observed motor interference was accomplished by bonding the negative lead from the motor directly to the case as shown by the dashed connection of Figure 3.2.1.1.8-B. Formerly this lead was bonded to an external structure which is not at the same potential as the case. Paragraph 1.8.1.1 emphasizes the fact that bonding to a structure does not, in itself, assure the existence of a true ground plane at most of the frequencies encountered in interference problems. As a result an impedance, Z, existed between these points as shown in Figure 3.2.1.1.8-B. This impedance is a part of the closed loop, A-B-C-D-A, through which interference currents generated by the motor flowed. The portion of the same loop, A-B-D-A, which was formed by connecting the negative lead from the motor terminal to the external ground, passed through strong interference fields. Interference voltages were induced in this loop as indicated by the equivalent interference generator in the figure. Since this impedance is also a portion of the loop, E-C-D-A-F, which contains the external power source, any interference voltage that appears across it causes interference currents to flow in the receivers which employ the same power source.

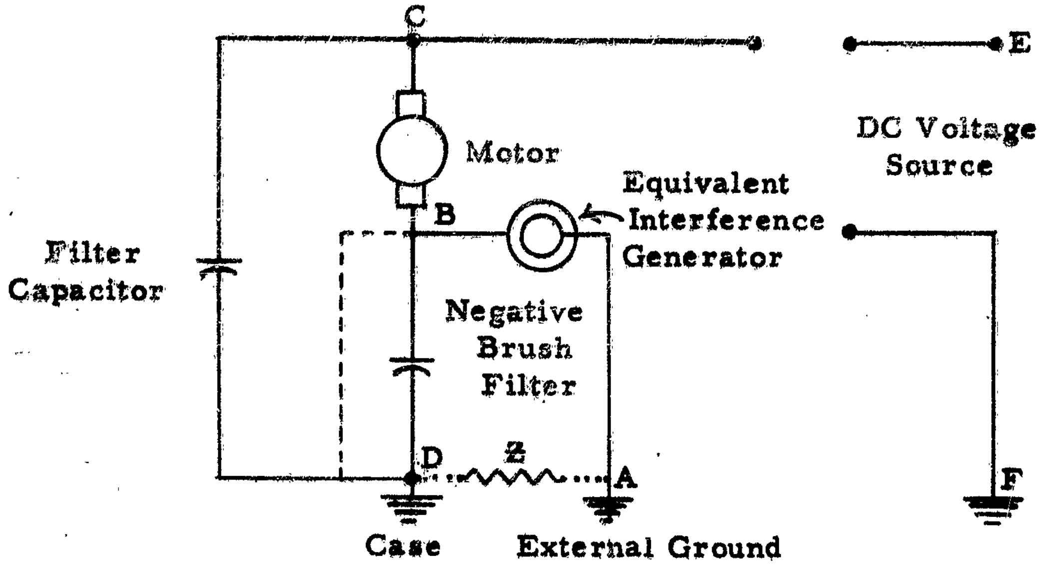


Fig. 3.2.1.1.8-B - Ambiguity of Ground Points

As a result of incorporating these modifications, it was found that in the region of 5 to 20 megacycles the radiated interference was reduced from a maximum value of 42 microvolts and an average value of approximately 20 microvolts down to a maximum value of 2 microvolts and an average value of approximately 1/3 microvolt. In the region of 1000 megacycles, no interference was measurable at any distance from the commutator end-bell after modification.

This example illustrates how oftentimes relatively minor changes in design can make a piece of electrical machinery interference-free. Nevertheless, the necessity of incorporating design changes no matter how minor is both costly and time-consuming. By following the principles of good design for interference-free operation, the need for making changes in production models in order to meet specified

interference limits could be avoided. No one design principle which pertains to adaquate bonding, filtering, or shielding should be neglected. Careful study of the unit being designed is required in order to provide means for attenuating or bottling up all possible conducted and radiated interference.

3.2.1.1.9 ALTERNATING-CURRENT MOTORS

Squirrel cage induction motors are not usually troublesome if they are carefully designed for minimum generation of harmonics. Motors with slip rings should be carefully designed because the installation of a filter is not usually practical due to the added weight. Alternating-current motors of the type that require commutators should be avoided altogether in aircraft.

3.2.1.1.10 INVERTERS

Inverters are needed to supply alternating current in aircraft whose primary power system consists of a set of direct-current generators. Rotary inverters are inherently sources of interference because they have commutators as well as slip rings, and, in addition, they produce a large number of harmonics of considerable amplitude both on the input and on the output sides.

Since basically a rotary inverter is a direct-current machine with added taps on the armature winding and slip rings connected to these taps, the design considerations for direct-current generators apply here in full force. But in addition to this, filters must be provided in all output and input leads. This imposes a particularly stringent requirement on the alternating current side because the required filters are heavy and bulky; but, for inverters of large size such installation is usually necessary and justified.

3.2.1.2 VIBRATORS

Vibrators are used to convert direct current into alternating current by means of vibrating contacts which alternately make and break the direct-current line. The wave forms of the resulting currents and voltages are more nearly rectangular than sinusoidal. They are therefore very rich in high harmonics and capable of producing a large amount of radio interference. If at all possible, the use of vibrators should be avoided and other means of converting direct into alternating current should be employed in aircraft. When it is necessary to use vibrators, complete shielding and extensive filtering must be employed to keep the interference from causing damage.

There is one particular application of vibrators where the radio interference is not objectionable and where they are being used extensively: Engine - starting vibrators operate only during the brief interval of starting the aircraft engine while the engine speed is not sufficient to allow the magneto to develop a high enough voltage. Once the motor has reached a predetermined speed, the starter switch is disengaged, which also cuts off the vibrator. Such an arrangement would not be suitable, for example, in aircraft operating on an aircraft carrier because radio interference even during the short starting period is not permissible in the vicinity of the many receivers on the carrier itself. But in land-based aircraft, starting vibrators are commonly used.

This starting vibrator, while not in itself a source of radio interference once the engines are running, causes radio interference problems of a different kind: The vibrator unit offers coupling paths through which the ignition interference fr . the magneto is introduced into the direct-current power system of the aircraft. The interference could be prevented from leaving the vibrator housing by filtering the power lead connecting the unit.

However, filters add weight and do not represent the most desirable approach for radio interference suppression in this case. It was found that appropriate design techniques applied to the vibrator unit itself accomplished the same purpose without the use of filters.

A sketch of a typical vibrator unit is given in Figure 3.2.1.2-A. There are two main component parts, an off-on relay and a vibrator, housed in a metal case.

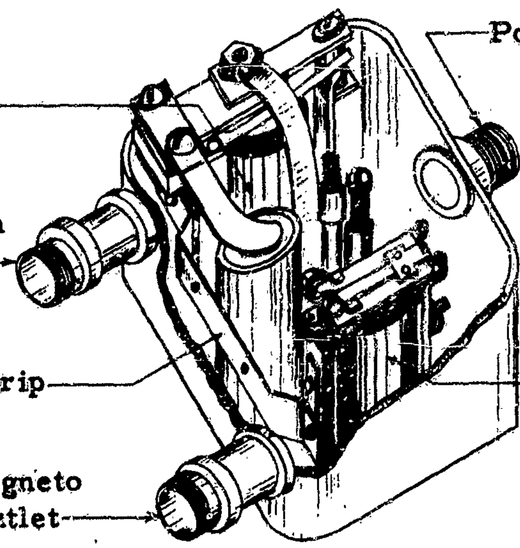


Fig. 3.2.1.2-A Aircraft Vibrator Unit, Original

The moving contact arms of relay and vibrator are attached to their metal frame and are electrically insulated from ground. A shorting strip connects the magneto and ignition switch outlets. A starting mesh-switch, not part of this unit, allows the relay contacts to close, which energizes the vibrator and supplies high surge currents to the magneto primary. This produces a high voltage in the secondary coil of the magneto, which supplies normal output to the spark plugs during the interval required by the engine to reach a minimum running speed. The primary winding of the magneto in series with a set of breaker points, where the opening and closing of the points create low tension pulses. In this way, steep-wave-front transients are produced, and the resulting radio-frequency energy is conducted back through the primary lead of the magneto to the vibrator.

The interference voltages gain admittance to the direct current aircraft wiring by the following coupling paths:

(a) Capacitive coupling across the open contacts of the off-on relay.

(b) Capacitive coupling between the shorting bar of the two outlets and the coil windings of the relay and vibrator.

(c) Capacitive coupling between the relay frame, which is directly connected to the magneto primary circuit and the shorting bar, and the winding.

(d) Electromagnetic radiation from the relay frame.

(e) Electromagnetic radiation from the shielding case, since the shield is broken by insulating gaskets on the base plate and cover.

A typical unit was subjected to five modifications, which resulted in a considerable reduction of the radio interference coupled to the direct-current power system. The modifications are the following:

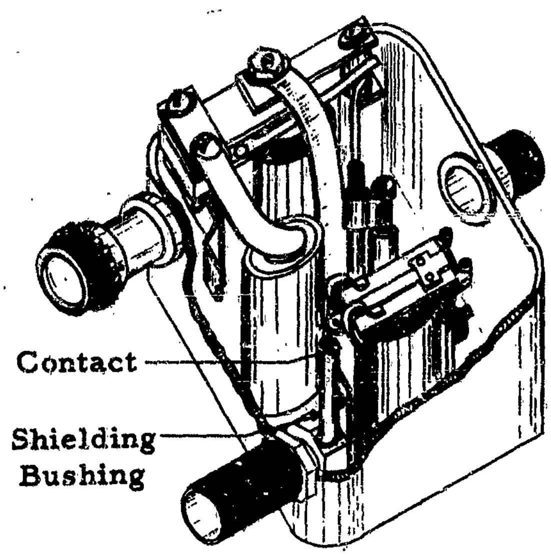
(a) The movable contact arm of the relay was tied to the vibrator instead of the magneto lead, as shown in Figure 3.2.1.2-D.

(b) The magneto lead, including the stationary contact point of the relay, was surrounded by a shielding bushing having only one small opening to allow the movable arm to make contact.

(c) The shorting bar was removed entirely. The connections to the ignition switch were no longer made through the vibrator unit.

(d) The vibrator coil was rewired to minimize capacitive coupling.

(e) A conductive gasket was installed to insure good electrical contact between cover and base plate.



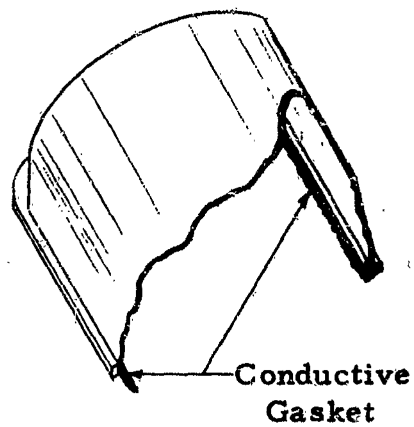


Fig. 3.2.1.2-B Modified Vibrator Unit

The vibrator unit featuring these modifications is shown in Figure 3.2.1.2-B. An additional modification is suggested (but was not carried through in the example

shown): The capacitive coupling across the open relay contacts may be eliminated by a grounding contact on the relay instead of the existing open position.

The schematic diagrams of the original and modified type are illustrated in Figures 3.2.1.2-C and 3.2.1.2-D for comparison purposes. Tests on the modified unit indicate an average reduction of 60 db in the frequency range of 0.25 to 18 megacycles, dropping to 40 db between 30 to 144 megacycles. Grounding the movable contact arm of the relay results in a reduction greater than 100 db between input and output circuits of the vibrator throughout the frequency range of 0.20 to 144 megacycles. The results of such modifications, without the application of filters, indicate that interference in aircraft equipment can be reduced to an acceptable level by the application of appropriate design techniques.

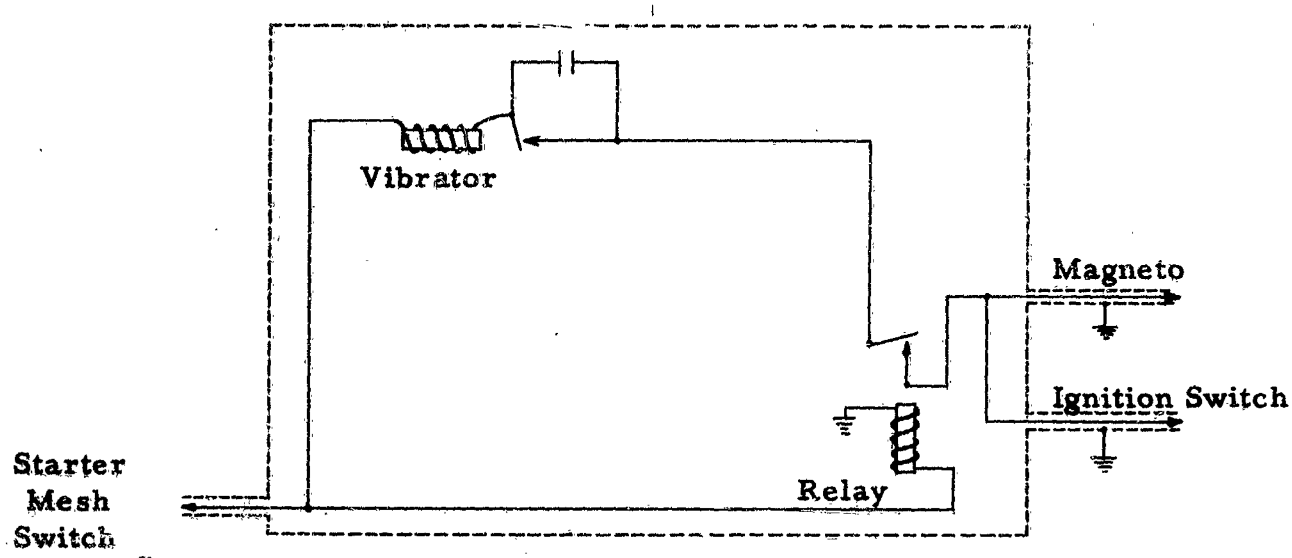


Fig. 3.2.1.2-C Schematic Diagram, Original Circuit

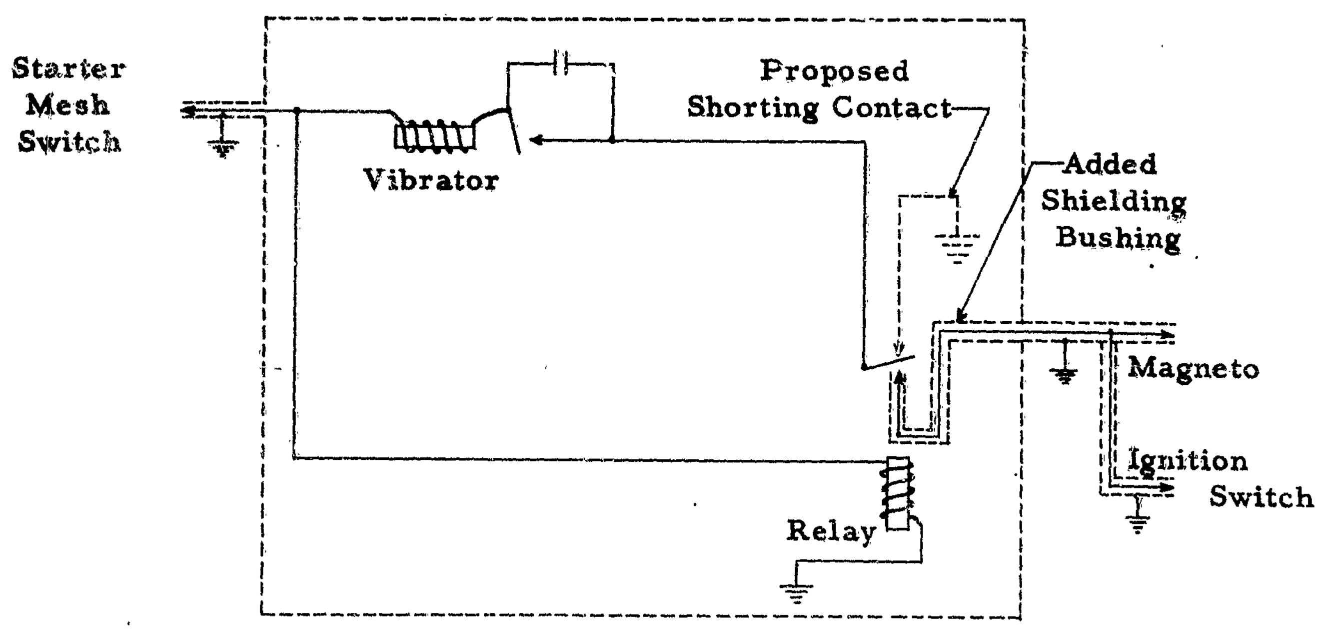


Fig. 3.2.1.2-D Schematic Diagram, Modified Circuit

3.2.2 ELECTRICAL OSCILLATIONS IN STATIONARY CIRCUITS

Radio interference generated by the operation of electronic equipment in an aircraft, generally originates in stationary resonant circuits. These include local oscillators, transmitters, and modulators, which perform functions essential to the operation of the airborne radio and radar equipment. The stationary sources of interference, like the motional sources, are scattered throughout the aircraft in such a fashion that each compartment and wing section may contain several pieces of electronic equipment. Figure 3.3-C shows the location in a typical heavy bomber of various electronic components and the routing of the interconnecting cables. This shows that any offending electronic component may be mounted sufficiently close to susceptible components or component wiring to introduce radio interference into these units and adversely affect their operation. Consequently, each electronic component must be designed so that electrical oscillations are confined to the unit itself and not permitted to enter other electronic components either directly or indirectly and cause radio interference.

Oscillations may occur in electrical systems whenever the passive circuit elements are connected so that electrical energy may be exchanged periodically between them. Such oscillations or resonant conditions can be excited by any appropriate frequency component of the applied signal. Oscillatory action of this nature would be self-sustaining if the resonant circuit contained no resistance. However, since every practical circuit contains resistance, the supplied energy is dissipated gradually, and the oscillation is damped. Therefore, the peak values of the current or voltage oscillations will decay exponentially unless the dissipated energy is returned to the system. Circuits containing inductive and capacitive parameters, capable of storing energy, will resonate if properly excited. Spurious oscillations as well as the desired resonant condition may be set up in an electrical system by such action. In any case the sustained oscillations must be either self-excited or externally excited. Any amplifying device, such as a vacuum tube, is capable of generating and sustaining oscillations. Due to the amplifying characteristic of the tube, power available in the output is much greater than the required input power. This permits a certain percentage of the output to be fed back into the input in the same phase as the input energy, and thus supply its own excitation. Any parallel-resonant circuit is an example of external excitation when a constantly varying electromotive force is applied across its terminals. The same type of tuned circuit experiences a series of transient oscillations when subjected to a sudden pulse of energy.

Regeneration is a first requisite for sustained oscillation in any circuit. The amount of regeneration of in-phase feedback must be sufficient to overcome the resistance losses present in the input circuit. Practical oscillators must be self-starting as well as self-sustaining. The self-starting feature is provided by a random fluctuation in voltage due to such causes as, thermal agitation in the circuit and tube, fluctuations in tube current produced by shot effect, and contact differences of potential. These random voltages cover the entire frequency spectrum, but the tuned circuit selects that particular frequency to which it has a natural response and amplifies it, by the regenerative process, so that an oscillating condition is maintained. Undesirable or "parasitic" oscillations result in many circuits when the conditions for resonance are present. A feedback path for sustaining "parasitic" oscillations can be supplied by the grid-to-plate capacitance of an ordinary amplifying tube, the capacitance between adjacent circuit wires, or the distributed capacitance of transformer

turns. Inductive coupling between transformers also becomes serious when there is a relatively high gain between them. It may be necessary to screen one or both transformers with metal of low reluctance. Proper wiring layout will normally prevent coupling between adjacent wires in components; otherwise shielded leads will be required. Negative feedback may be employed to neutralize the effects of tube capacitance coupling.

3.2.2.1 LOCAL OSCILLATORS

In superheterodyne receivers the local oscillator circuits serve a useful purpose in supplying an output whose frequency differs from a received signal frequency by a constant difference. The oscillator energy must at the same time be prevented from coupling through any path leading to the antenna or chassis and risking radiation which will interfere with the normal operation of adjacent electronic equipment. This is discussed under Paragraph 3.4.4.

3.2.2.2 TRANSMITTERS

Radio transmitters are generators of radio frequency energy which is controlled by the intelligence to be transmitted. The very heart of such a generator is the oscillator circuit whose frequency of operation must be highly stable for the usual transmitter applications. Accordingly under most conditions the frequency stability of a crystal oscillator is desired. However, the frequency range in which crystals may be employed is limited, and at the higher radio frequencies harmonic generators must be used. An important exception exists in microwave systems where techniques of operation have not been fully developed and the spectrum is not at all crowded, and frequency control is often obtained by the use of cavity resonators. To accomplish their purpose with a high quality of transmission, radio transmitters must be free from harmonic radiation, spurious sidebands, distortion, and hum. Radiation of harmonics is particularly troublesome in high-powered transmitters. One percent of second harmonic radiation in a 50-kw transmitter corresponds to a 5 watt power signal level, and can readily produce an interfering signal over considerable area.

Negative feedback is frequently used in transmitters. The schematic diagram shown in Figure 3.2.2.2-A illustrates the operation of a feedback amplifier.

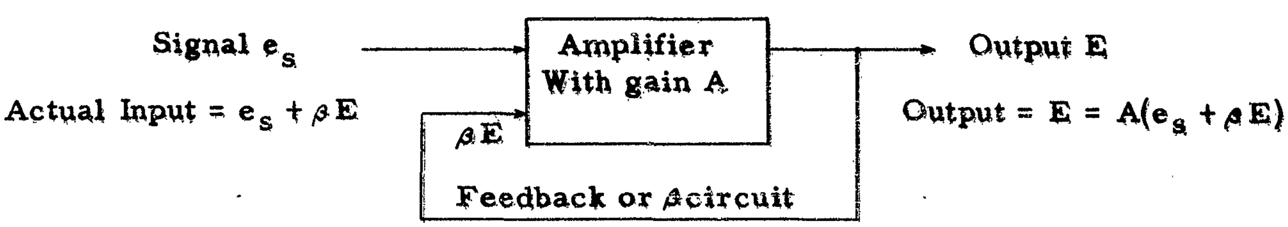


Fig. 3.2.2.2-A Feedback Amplifier Schematic

The quantity A represents the amplifier gain, and e_s the supplied signal. A fractional part, βE, of the output E is added to the external input, so that the actual input consists of the signal e_s plus this fractional part βE. The total input, multiplied by the gain, must equal the output. Thus, $E = A(e_s + \beta E)$. Representing the gain

as the ratio of output to external input, it is simple to derive a gain formula.

$$\begin{aligned} E/e_s &= \text{actual gain} \\ E &= A(e_s + \beta E) \\ E &= Ae_s + A\beta E \\ Ae_s &= E - A\beta E \\ \therefore E/e_s &= A/(1-A\beta) = \text{gain with feedback.} \end{aligned} \tag{3-29}$$

If the feedback opposes the input signal, then β is considered negative. The quantity $A\beta$ is called the feedback factor, and if this quantity is much larger than unity, then the gain is reduced to $-1/\beta$. Thus when the feedback is large, the effective amplification depends only on the value of β, and is practically independent of the actual gain produced by the amplifier. If the feedback circuit employs a resistance network, the gain is almost independent of frequency but has phase shift of approximately 0° or 180°. If it is desired to have amplification vary with frequency, then the feedback circuit (β circuit) can be designed to have the desired transmission-loss characteristic.

Negative feedback causes a reduction in amplitude distortion since some of the distortion is fed back to the input through the feedback circuit and reamplified in such phase as to cancel out most of the original distortion. If D indicates distortion in the output and d the distortion generated in the amplifier, then

$$D = d/(1-A\beta) \tag{3-30}$$

A large feedback factor, (recall that β is negative for opposing phase), will greatly reduce distortion in the output.

Feedback will modify the signal-to-interference ratio by the following relationship. Signal-to-interference ratio with feedback/signal-to-interference ratio without feedback = $A_f/A_o(1-A\beta)$. A_f represents the amplification taking place between the point where the interference is introduced and the output, with feedback. A_o represents the same amplification without feedback. The equation is based on the assumption that it is the same interference in each case and that the output voltages are the same. Analysis shows that feedback will reduce interference introduced in the high level stages of the amplifier, as for example a poorly filtered power supply in the plate circuit of the final tube. It will not aid in reducing interference entering the low-level stages, as microphonics or induced hum, since the feedback affects the interference and normal signal output to about the same extent.

The preceding paragraphs serve as background for appreciating some of the benefits of negative feedback. However, oscillations can result from such feedback because of accompanying phase shift. In the normal range of frequencies, circuit arrangements are such that feedback is negative. At the very low and very high frequencies, the amplifier stages produce phase shifts sufficient to cause the feedback factor, $A\beta$, to change from negative to positive. Oscillations are not usually encountered in two-stage arrangements unless the feedback factor is made large. Where there are more than two stages, however, oscillations tend to take place even with a moderate amount of feedback.

Phase shift depends upon variation in the amplitude of transmission with frequency and its polarity on the sign of the slope of the transmission characteristics.

Where the transmission is constant with frequency, that is a flat response, there is no phase shift. If the amplitude of transmission, "a", has a constant variation, then the phase shift is directly proportional to the slope of the amplitude characteristic. Expressed as an equation -

$$\text{Phase shift in radians} = \frac{\pi}{12} \times \frac{da}{du} \qquad (3\text{-}31)$$

where da/du represents variation in the amplitude of transmission expressed in decibels change in transmission for an octave change in frequency. In designing a feedback system to avoid oscillations it is only necessary to consider the way in which the amplitude of transmission varies with frequency. Thus if the feedback factor, $A\beta$, has shifted in phase by 180° and becomes positive, the magnitude of $A\beta$ must have decreased to less than unity to avoid oscillations.

In frequency-modulated transmitters negative feedback will give the same benefits as in the amplitude-modulated system. The problem of preventing oscillations will also be the same and therefore subject to the same design considerations previously discussed.

3.2.2.3 MODULATORS

Radar modulators are capable of producing large amounts of interference. The problem of reducing the interference to meet required specifications can and must be solved by proper design and internal shielding.

Measurements were made on a line type modulator to determine the best method for reducing interference conducted along the power line. A single copper sheet between primary and secondary windings served as an electrostatic shield for the power and filament transformers. Primary leads were placed as far from the thyratron as possible and at right angles to other wiring. The primary power leads were shielded with copper tubing from the transformers to the point where they left the modulator case, and this shielding was grounded. Partition shielding was used to separate the rectifier from the thyratron. These features are illustrated in Figure 3.2.2.3-A.

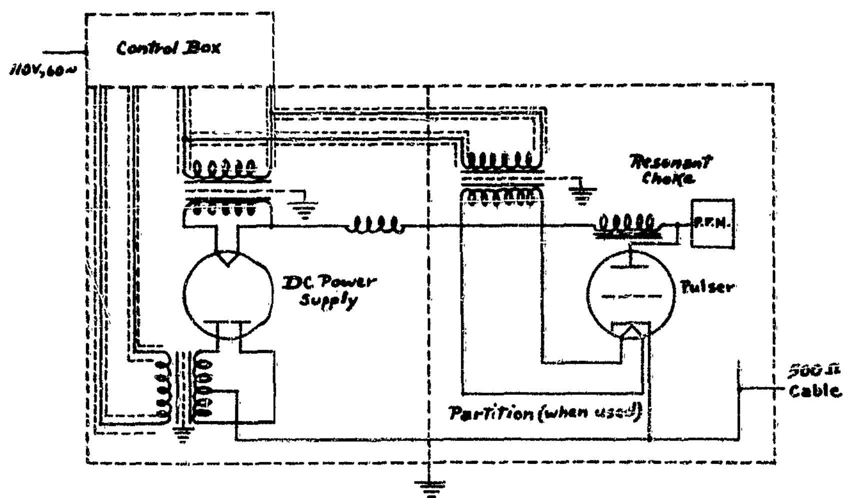


Fig. 3.2.2.3-A Radar Modulator and Pulse Lead Shielding

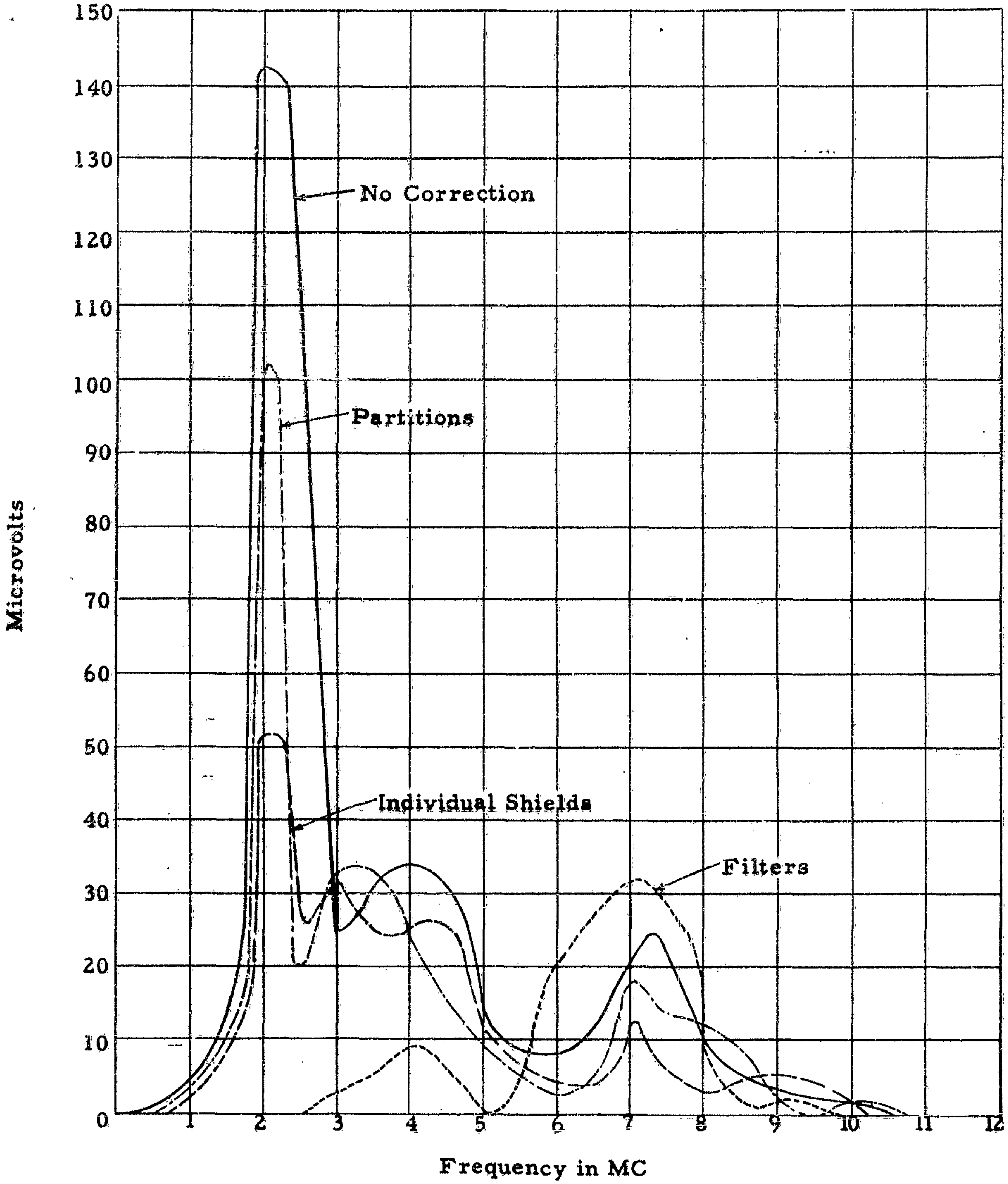


Fig. 3.2.2.3-B Interference Plotted Against Frequency

Interference was reduced to less than 50 microvolts over the entire spectrum, from 0.2 to 20 megacycles, without adding filters. When used, filters offered additional interference suppression from 0 to 5.5 mc, but the difference was minor above 5.5 mc, and therefore without justification in view of their additional weight and size. Individual shields were also used in place of the partition, and offered considerably

greater reduction in the region of two megacycles. The relative merits of these design features are shown in the graph of Figure 3.2.2.3-B.

3.2.2.4 TRANSFORMERS

Iron core transformers have a non-linear relationship existing between flux and magnetizing force which produces harmonics in the lower and upper regions of operation. With high flux densities harmonic generation results from core saturation. The permeability of the iron core is not constant and hence the inductance is not constant. This causes the input impedance of the transformer to vary in a non-linear fashion which produces currents not present in the original input as discussed in Paragraph 1.3.1.3. At low values of flux density, harmonics are also produced, but they are caused by the variation of that part of the total resistive component of the input impedance which represents the hysteresis loss. As this portion of the resistive component varies, so will the power loss vary, increasing in a complex manner with the flux density. The higher order harmonics are particularly troublesome since they may lead to resonance conditions between the secondary inductance and the distributed capacitance. The energy, once developed, may readily find a coupling path to nearby receivers through power leads or even produce radiation strong enough to interfere with adjacent equipment. Harmonics developed within a receiver by the power transformer may be coupled to the antenna with consequent radiation, through the capacity existing between primary and secondary. This latter effect, however, is readily cured by the use of a Faraday shield. A variable permeability at high flux densities may be avoided within extended limits by using better grade cores suited to a particular purpose. Permalloy, hipernik, mumetal, perminvar, among others are characterized by high maximum permeability. There is far less harmonic generation when using permalloy than when using silicon steel, and perminvar is many times better than permalloy.

The designer of aircraft electronic equipment, mindful of the severity of the interference problem, must carefully select an appropriate transformer suited to a particular purpose wherever iron-core transformers are to be used.

3.2.2.5 PARASITIC OSCILLATIONS

Oscillations which occur at other than a desired frequency, or outside a tank circuit, are called parasitic oscillations. They may take place in oscillators as well as in ordinary power amplifiers, and the energy they represent is capable of reducing the normal output at the operating frequency to a small fraction of its value. These spurious frequencies give rise to distortion in linear amplifiers and modulators, and may produce spurious side bands, cause flashovers, and other undesirable effects.

Parasitics may be of higher or lower frequency than the normal operating frequency of the amplifier or oscillator. When a circuit possesses sufficient energy-storage capabilities and enough feedback of the proper phase, it will oscillate, and the effect is normally super-imposed on the output of the amplifier or oscillator.

High frequency parasitic oscillations, usually above 30 megacycles, may exist in tuned radio-frequency electronic circuits. The circuits give rise to parasitic oscillations due to the lead inductance between tube and tank circuit and the interelectrode capacities of the tube. When large tubes are used, the long leads and rather

large interelectrode capacities, as well as high transconductance of a given tube, increase the possibility of parasitic oscillations. Ordinary triodes have a feedback path through the grid-plate capacitance. In pentodes there is coupling at high frequencies since the screen and suppressor grids are no longer at zero potential because of the ground lead inductance. The coupling capacities are from plate to suppressor, and from grid to cathode. Part of the lead inductance is due to the internal wiring of suppressor and cathode, as shown in Figure 3.2.2.5-A, and the remainder due to the lead wires of the by-pass capacitor to ground. The capacitor itself, at relatively high frequencies, is like a short circuit. This analysis will explain why tubes designed for ultra high frequency applications have a separate suppressor grid pin on the tube base. This should be directly connected to the chassis and not to the cathode.

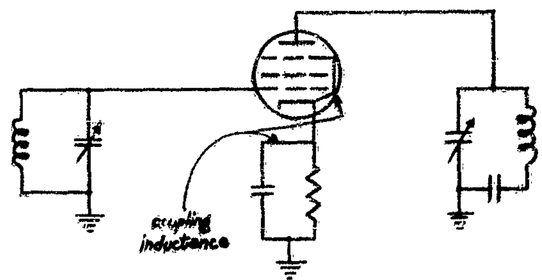


Fig. 3.2.2.5-A Schematic of Lead Inductance at High Frequencies

For a better understanding of parasitic oscillations and the means of preventing them at frequencies well above the normal range of operation, consider the class C amplifier illustrated in Figure 3.2.2.5-B.

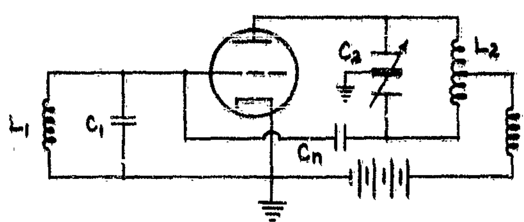


Fig. 3.2.2.5-B Conventional Class C Amplifier

At high frequencies the capacitances C_1 and C_2 of the tank circuits are practically short circuits, while the tank inductances, L_1 and L_2 may be treated as open circuits due to the high impedances at the frequencies involved. Under these conditions, the circuit reduces to a tuned-grid-tuned-plate oscillator type, as shown in Figure 3.2.2.5-C. The grid and plate tuning capacities are supplied by the interelectrode capacitances of the tube, and the inductances, L_g and L_p, by the lead inductances between electrodes. By comparing this figure with Figure 3.2.2.5-B, it is noted that the neutralizing capacity is not effective since it forms no part of the parasitic oscillatory circuit.

The parasitic oscillations may be prevented by inserting a small resistance, about 1 to 25 ohms, in series with the grid or plate lead. It is preferable to insert it in the plate lead since it then affects the parasitic current directly, but is not in series with the main oscillating circuit. Detuning processes are effective in eliminating parasitics. The resonant frequency of the grid circuit may be increased, or that of the plate decreased, which causes the plate circuit to offer a capacitive reactance and thus introduces positive resistance into the grid circuit. The detuning process may also be accomplished by shortening the grid leads to decrease their inductance and lengthening the plate leads, or inserting a small choke in the plate lead next to the tube. The resonant frequency of the plate circuit being much lower than the grid circuit, oscillations cannot occur.

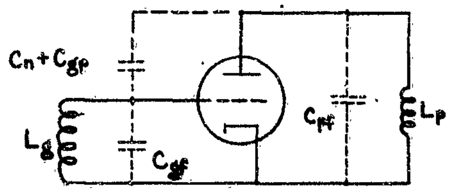


Fig. 3.2.2.5-C Class C Amplifier, Equivalent Circuit

Low frequency parasitics occur where radio-frequency chokes are used in series with the DC supply to both plate and grid, as shown in Figure 3.2.2.5-D, while the equivalent circuit is shown in Figure 3.2.2.5-E. At frequencies well be-

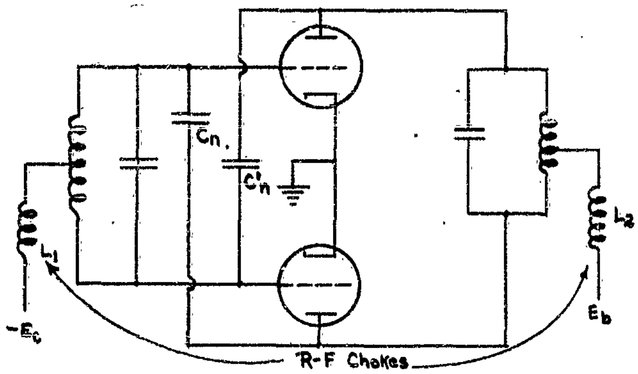


Fig. 3.2.2.5-D
Amplifier Circuit

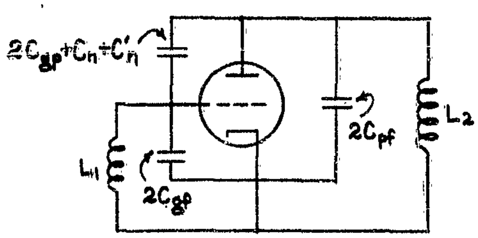


Fig. 3.2.2.5-E
Equivalent Parasitic Circuit

low the operating frequency the tank circuits' inductances are effectively short circuits, and the equivalent tuned-grid-tuned-plate circuit employs the chokes, L_1 and L_2, as inductances and the interelectrode capacitances of the tubes as tuning capacitances. Again the neutralizing condensers are not effective in preventing coupling, since the tubes act in parallel for the parasitic action, rather than push-pull. Chokes are to be avoided, if possible. If, however, it is necessary to use them, oscillations will be prevented by a selection of chokes such that the resonant frequency of the grid circuit is higher than that of the plate.

Parasitic oscillations may occur in the grid circuit of a vacuum tube. Throughout a given voltage range, grid current may decrease as grid voltage increases because of secondary emission. This indicates a negative grid resistance over that range and may result in parasitic oscillations. Besides the possibility of undesirable oscillations, severe distortion would be present in the output, and operation within the voltage range where this action occurs is to be entirely avoided by proper adjustment of the plate voltage and the grid bias.

Pentagrid and triode-hexode converter tubes, which combine the functions of oscillator and mixer tubes, are characterized by a type of interaction which results from coupling between the signal grid and virtual cathode in the vicinity of the signal grid. The virtual cathode pulsates at the oscillator frequency and thus induces currents at the same frequency in the signal grid circuit. This effect increases with frequency. At high frequencies, the tuned circuit of the signal-input grid has a resonant frequency that may differ from the oscillator frequency by only a small amount. Thus considerable impedance is offered to the induced current of the local oscillator frequency. The result is that the oscillator voltage developed on the signal grid causes the output to drop and to become relatively independent of the input tuned circuit. This space-charge coupling can be effectively neutralized by inserting a capacitance in series with a small resistance between signal and oscillator grids. This arrangement is indicated, by dotted line, in Figure 3.2.2.5-F.

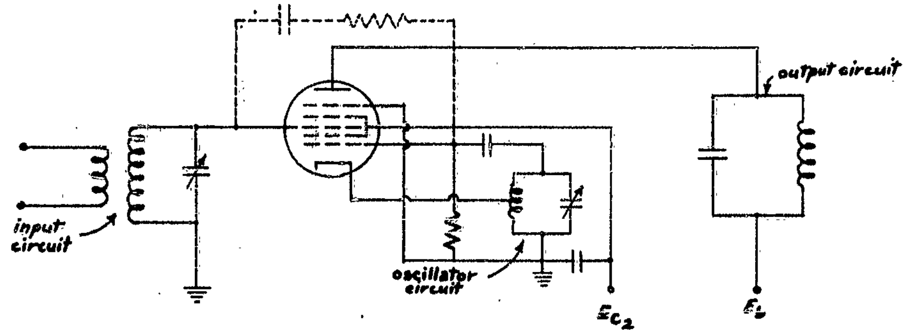


Fig. 3.2.2.5-F Neutralization of Space-Charge Coupling

3.2.2.6 TRANSIENTS IN RESONANT CIRCUITS

The circuits previously mentioned in Paragraph 3.2.2.5, which were capable of setting up parasitic oscillations, may just as easily set up transient oscillations. The parasites represent the condition of sustained oscillations, while the transients are damped oscillations decaying with time. Transients may be produced in any circuit when it is disturbed by any sudden electrical change, such as a pulse of energy, or the discharge of a capacitor. Transient oscillations may produce interference in a receiver output such as reduction of the output and introduction of distortion and spurious frequencies. A group of damped oscillations, repeated periodically may produce interference in the output at the repetition frequency. If only one or two oscillations of appreciable intensity are present, or if single pulses of negligible width act on a resonant circuit, they may produce disturbances in the output due to their broad energy spectra as explained in Paragraphs 1.2 and 1.7.2.

Oscillations produced in coupled circuits are more complex than a simple damped train of oscillations in a single circuit. An analysis of transients in electrical

systems results in a mathematical expression of the voltage-time, current-time, and charge-time relationships of a system. The equations involved indicate the nature of the oscillations, and how they may be prevented. Applying Kirchoff's voltage law to the series LCR circuit shown in Figure 3.2.2.6 yields the equation

$$E - iR - L\frac{di}{dt} - \frac{1}{c}\int i\,dt = 0 \qquad (3\text{-}32)$$

An analysis of this equation was made in Paragraph 1.8.1.2 and it was shown there that critical damping exists for the condition $R = 2\sqrt{L/C}$. If R is less than $2\sqrt{L/C}$, the circuit is underdamped, the dissipation is small, and the circuit will be subject to damped oscillations. If R is greater than $2\sqrt{L/C}$, it is overdamped and oscillations are prevented. The circuit shown is used chiefly for the purpose of analysis, and the oscillations could result from the actual closing of a switch, or from a pulse of energy inductively coupled to the circuit inductance.

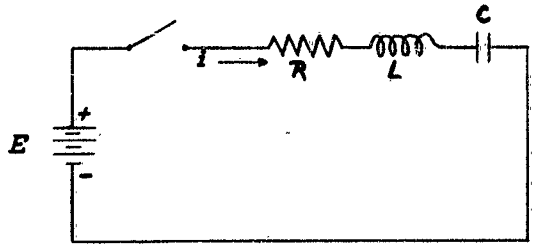


Fig. 3.2.2.6 Series LCR Circuit

3.2.3 ARCING

Arcing as a source of radio interference is discussed in Paragraph 1.3.2.4. Most arcs that are troublesome in electrical or electronic equipment occur in connection with switching processes. Switching transients cause radio interference even in the absence of arcing, but usually the interference is increased by a large factor when arcs are present. Arcing is most severe in circuits having high inductance because it is the magnetic energy stored in the inductance that must be dissipated in the arc. Basic arc-suppression techniques are discussed in Paragraph 1.8.1.2.

3.2.3.1 IGNITION SPARKS

Present aircraft engine ignition systems are good examples of radio interference generators of the type discussed above because of the steep-wave transients that ensue immediately after the firing of each spark plug. Figure 3.2.3.1-A shows the wave shape of an ignition pulse when the output of a magneto is connected to a typical ignition system. This figure shows that the ignition pulse consists of a fundamental and a series of high frequency harmonics that are the cause of radio interference originating in the ignition system.

This interference can be prevented from being radiated or coupled into other aircraft wiring if the entire system is encased in a continuous metallic shield which is adequately bonded to the aircraft structure. The use of filters is usually not satisfactory because any filter sufficiently efficient to remove all high frequency components

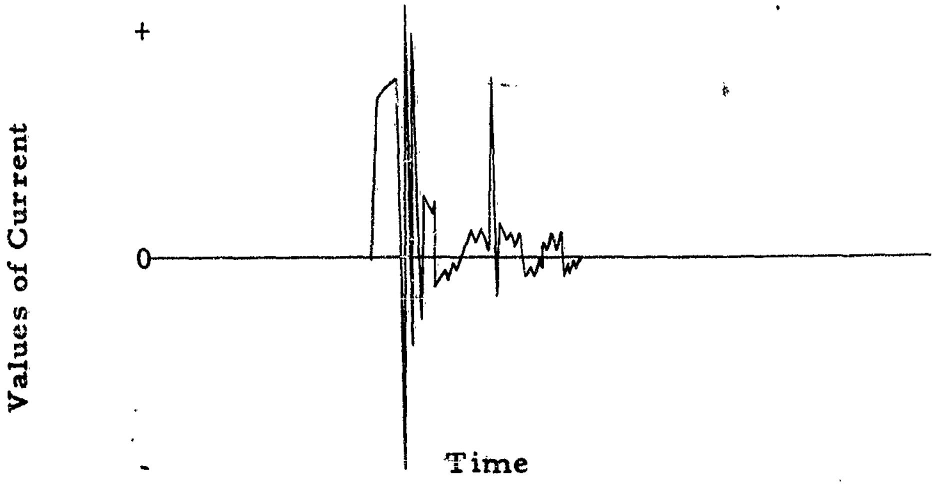


Fig. 3.2.3.1-A Wave Shape of an Ignition Pulse

of an ignition pulse would also destroy the characteristic wave shape which is essential to the correct functioning of the ignition system. Spark plugs, which produce the ignition spark by using the power and voltage developed by the magneto, must also be shielded to prevent the radiation of radio interference energy. Figure 3.2.3.1-B shows a typical spark plug, threaded into a recessed well in the cylinder head, as well as the finned metallic enclosure which is employed to shield and cool the plug.

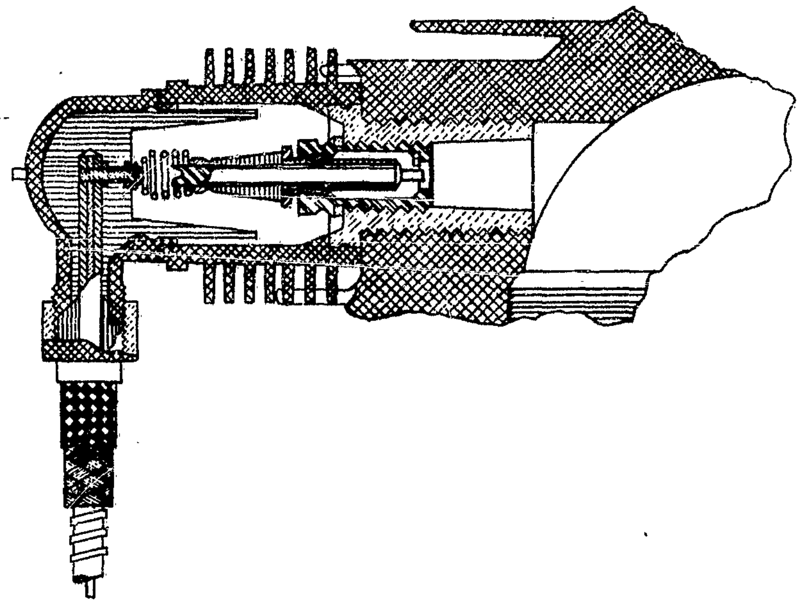


Fig. 3.2.3.1-B Spark-Plug Cooler and Radio Shield

Suppressors, i.e., impedance elements placed in series with the high tension lead of the distributor, are also employed to decrease the level of ignition interference in aircraft. The purpose of suppressors is threefold:

(a) To reduce the energy of the ignition impulse to such a value that spark plug erosion may be minimized.

(b) To render the circuit non-oscillatory.

(c) To reduce the steepness of the pulse wave front.

It is necessary to reduce the steepness of the wave front because even single periodic pulses are capable of setting up oscillations in the resonant circuits of adjacent receivers if the wave front is sufficiently steep. The steepness of the wave front is lowered by suppressors because the added resistance increases the time required for the current to rise from zero to its maximum value or to fall to zero from its maximum value.

Suppressors, however, have not come into general usage in aircraft for the following reasons:

(a) Suppressors having a resistance of the order of magnitude 10,000 to 20,000 ohms have been found to reduce spark plug erosion and eliminate radio interference, but they have caused excessive fouling of the spark plug points and insulators. This results in the malfunctioning of the ignition system.

(b) Suppressors having a resistance of the order of magnitude 1000 ohms have met with limited success, but the difficulty in obtaining and maintaining the necessary permanent resistance-temperature characteristics have prevented further development.

3.2.3.2 T-R BOXES

The transmit-receive box is a cavity-gas switch used in a microwave radar set employing a single antenna for transmission and reception. It prevents the transmitted pulse from entering the receiver and does not interfere with the reception of the reflected pulse. The box contains a tube, a resonant cavity which can be tuned, and provision for coupling the input and output circuits to the cavity. Two conical metallic electrodes separated by a short distance are enclosed in the tube, together with a slight amount of water vapor to improve the recovery time. The transmitter pulse causes a spark discharge in the tube which detunes the resonant cavity. This introduces a high degree of attenuation between transmitter and receiver circuits. The cavity in a discharging state is a means of rejecting the flow of radio frequency energy, whereas in the non-discharging state, there is a good match between input and output with very little reduction in delivered power. The various elements, with their connections, are illustrated in Figure 3.2.3.2-A.

The T-R box must be well shielded in order to prevent the interference due to the arc from affecting adjacent equipments. Energy entering the receiver must pass through the T-R box, and during the interval of a transmitted pulse, a small amount of transmitted energy will enter the receiver. This is referred to as leakage power. The leakage power consists of three components; (1) the "spike" of energy, (2) the flat power, and (3) direct coupling power. These terms are explained in Figure 3.2.3.2-B, which shows a typical leakage power pulse together with an idealized pulse for comparison.

The energy contained in the spike contributes to converter crystal failure, and to a far smaller extent so does the power in the flat section of the pulse. The time

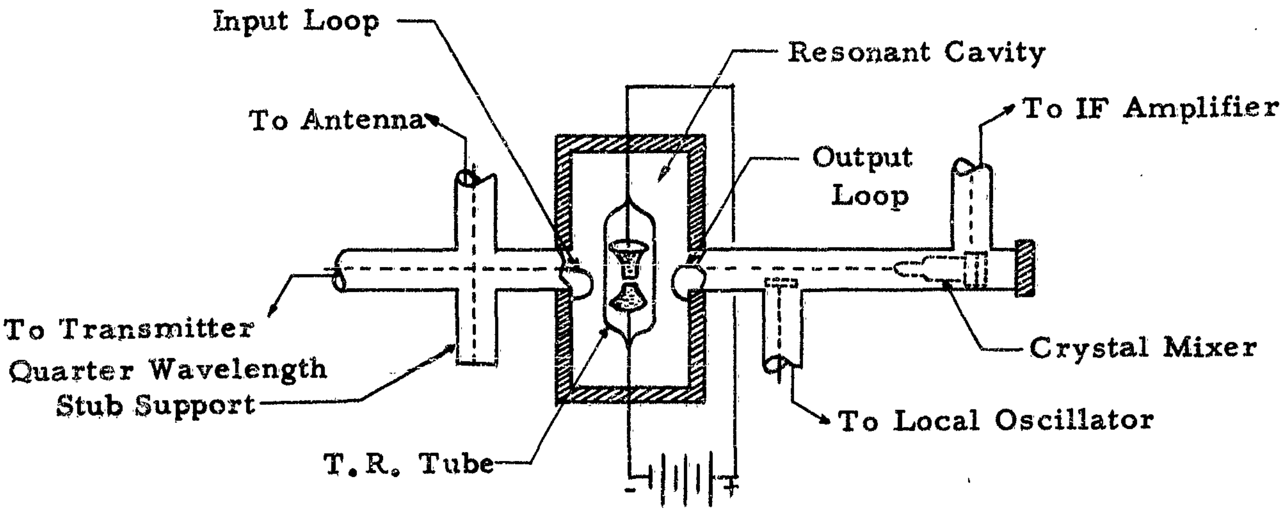


Fig. 3.2.3.2-A T-R Box and Crystal Mixer

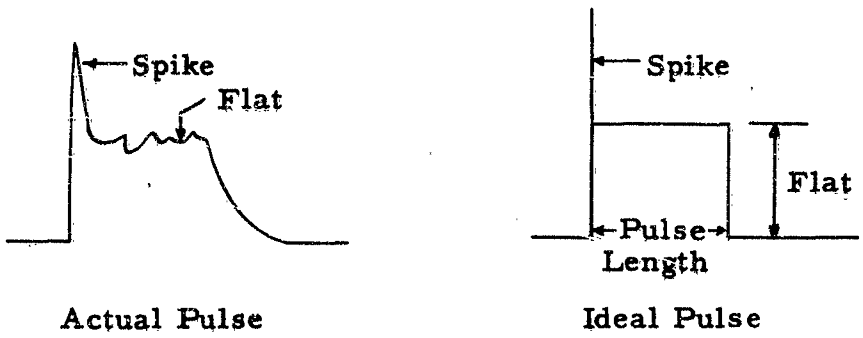


Fig. 3.2.3.2-B The Shape of the Leakage Power Pulse

interval of the spike has been estimated at approximately 1/1000 of a microsecond, and the energy content is mostly dependent on the steepness of the transmitted pulse wave front and the repetition rate. At low repetition rates, (less than 1000 pulses per second), the spike energy may be reduced by a direct-current glow discharge near the radio frequency gap. A "keep-alive" electrode is supplied in all standard T-R tubes which provides a continuous supply of ions and free electrons to help establish the desired conditions in the radio-frequency discharge path. However, oscillations may result due to the negative-resistance characteristics of the low current discharge. This produces a cyclic variation in the number of free electrons and ions in the gap and causes the spike energy to fluctuate. The mean free path of an electron is, in general, of the same order of magnitude as the distance between electrodes, but very few electrons reach the electrodes because of the very rapid variations in the radio-frequency field. Therefore, electrons oscillate back and forth, losing energy to the neutral gas molecules and to positive ions through occasional collisions. A limiting resistance mounted close to the cap of the "keep-alive" electrode will minimize the effects of these undesirable oscillations. If the oscillations persist, it is evidence of tube failure or a supply voltage that is too low. The auxiliary discharge current may be increased, but the T-R tube life would be correspondingly reduced.

When the radar set is first turned on, there are no residual ions in the discharge gap, and during the time of the first few pulses the spike may have a dangerously high value. A "crystal gate" is usually provided to isolate the crystal from the T-R box until stable transmitting conditions have been reached, and until the T-R tube discharge has been established. Another important function of the gate is to offer protection to the crystal, when the radar is idle, from damage by the radiated energy of other radars operating nearby.

The flat portion of leakage power indicated in Figure 3.2.3.2-B is, in general, proportional to the spacing of the gap. Most crystals withstand flat power levels very well. Direct coupling, however, which is directly proportional to transmitter power and is a component of the flat portion, will result in damage to the crystal under improper operating conditions. This is so when the magnetron develops an appreciable amount of power at frequencies other than the frequency of normal oper-~~ation. If they are in the vicinity of the resonant frequencies for these direct coupling~~ modes in the T-R box, they may easily be transmitted with little attenuation. This is a threat to efficient operation in very-high-power systems.

3.2.3.3 RELAYS

High speed oscillographic measurements of radio interference produced by relays lead to the following conclusions:

(a) The coil of the relay can be replaced by a capacitor having a capacitance equal to the distributed capacitance of the coil without altering the shape of the current transient wave during the first few microseconds after power is supplied to the circuit, as shown in Figures 3.2.3.3-A and B. This indicates that the distributed capacitance effectively "shorts out" the coil during this short interval and is responsible for the generation of interference transients.

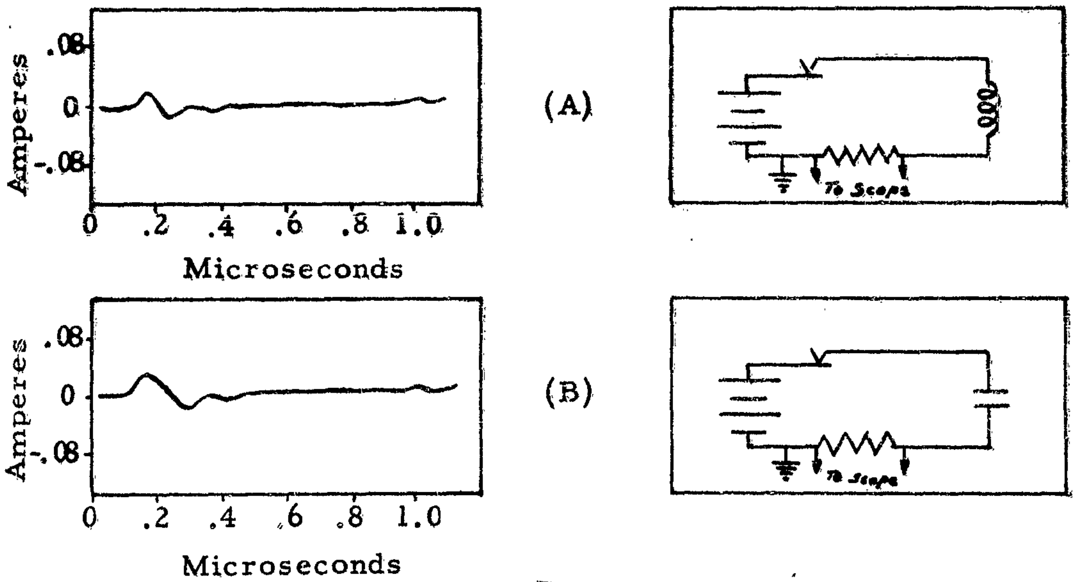
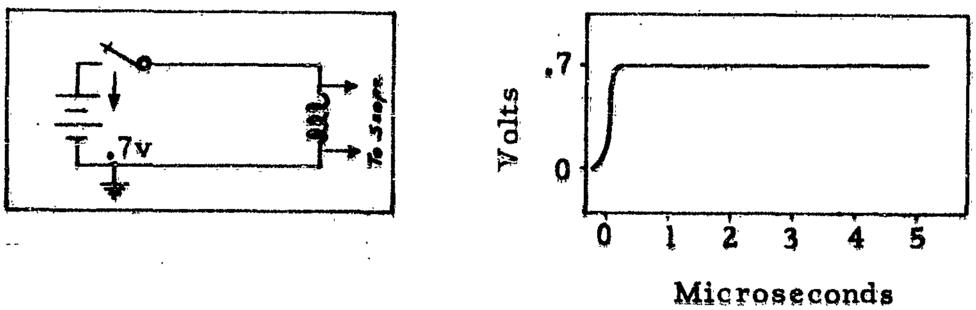


Fig. 3.2.3.3-A and B Equivalence of Relay Coil to Capacitance in Initial Closure Current Transient

(b) Typical relay coils exhibit a high ratio of inductance to distributed capacitance, which results in high amplitude voltage surges with steep wave fronts caused by the collapse of the magnetic field about the relay coil when the current is interrupted. Figures 3.2.3.3-C and D show that the voltage across the coil rises quickly to the supply voltage of 0.7 volts DC when the circuit is closed, but on the break the potential rises to a value of approximately 100 times the supply voltage in about three microseconds and then decays to a zero value at a rate determined by the inductance, distributed capacitance, and resistance of the winding. It should be emphasized that this voltage surge possesses a steep wave front which is capable of producing violent shock excitations in receivers tunable over a wide range.

(c) Switching units, with the exception of mercury switches, display mechanical bounce or chatter which causes repetitive closures and interruptions of the current when the switch is closed. The long duration sweep shown in Figure 3.2.3.3-E shows the effect of the bouncing switch contacts when the circuit is made. The high amplitude voltage surges shown in the figure are evidence that the points remain in contact sufficiently long to establish an appreciable current in the coil. These transients developed at the make of the circuit are of greater duration and severity than those developed at the break of the circuit.

(C)



(D)

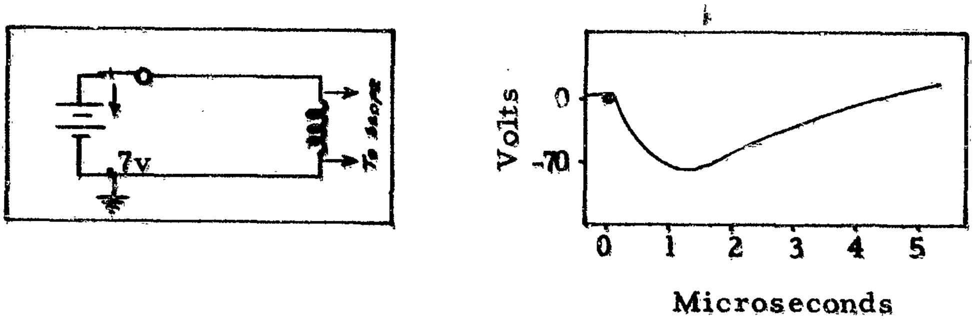


Fig. 3.2.3.3-C and D Make and Break Voltages Across a Relay Coil

(d) In addition to the transients due to mechanical bouncing, there occur rapid closures and interruptions of the circuit. These are at a faster rate than those due to the mechanical bouncing of the relay contacts at the make of the circuit as described above. This is shown in Figure 3.2.3.3-F. As the contacts move outward, the contact area for the flow of current decreases resulting in local heating, which causes the contacts to pit and sputter until the circuit is broken. The amplitude of the resultant induced voltage is sufficiently high to imitate "cold" emission from the projecting area of the relay contacts. This is accompanied by local heating, which

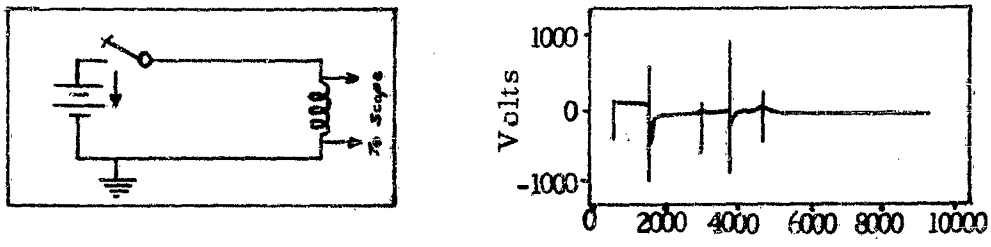


Fig. 3.2.3.3-E Bouncing Transients in Closing a Relay

causes the contact material to melt and neck out until the circuit is again closed. This process repeats at an exceedingly rapid rate until the relay contacts are separated far enough to prevent the voltage gradient from rising to the value necessary for cold emission. These closures and interruptions of the circuit are also responsible for the generation of steep wave forms, which cause radio interference in adjacent electronic circuits.

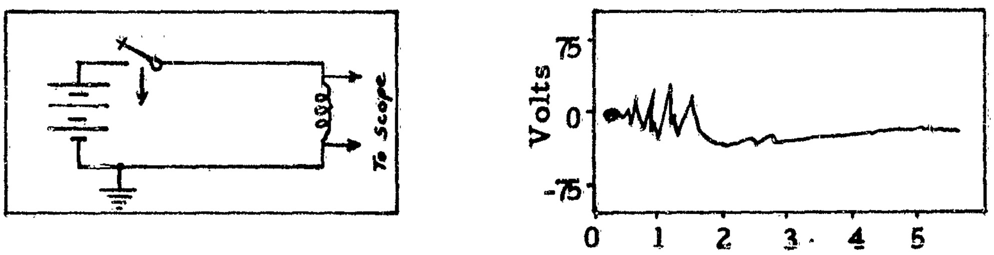


Fig. 3.2.3.3-F Pitting Transients in Opening a Relay

A reduction of the interference is obtained by enclosing the offending relay and its associated circuit within a metallic shield. However, direct oscillographic measurements show that singly shielded conductors are incapable of completely eliminating radiations from the central conductor of a coaxial cable. A typical measurement shows that the ratio of pick-up voltage of an adjacent external pick-up wire to the central conductor of a coaxial cable is approximately 1 to 1000. Therefore, it is frequently necessary to resort to multiple shielding of conductors along which surges of extreme sharpness are propagated if effective interference reduction is to be obtained.

The use of low pass filters is probably the most practical means of suppressing the interference developed by relays or other devices which develop similar steep wave transients. Studies of low-pass networks reveal that a series-inductance, shunt-capacitance network transmits the least high frequency energy. Filters of this type, however, fail to effect complete surge suppression because of the distributed capacitance of the inductance and the inductance that is inherently present in the capacitor leads. This suggests that the most effective low-pass filter should consist of an inductance with the least possible distributed capacitance in conjunction

with a feed-through capacitor. Capacitors of this type are described in Paragraph 3.1.1.5. Furthermore, losses may be introduced into the filter to dampen the low frequency oscillations that may appear in this system.

A resistor in series with a capacitor connected across the contacts is another method used for the suppression of interference. A capacitor should never be connected across the contacts without including a series resistance because the discharge of the condenser when the contacts are closed can cause a heavy surge if not controlled by the resistor. Refer to Paragraph 1.8.1.2 for a detailed discussion of this method of arc prevention.

3.2.3.4 POOR BONDING

One of the purposes of bonding, as explained in Paragraph 1.8.1.1, is to prevent arcing between adjoining metal parts. Bonding techniques are covered fully in the Military Specification on Electrical Bonding for Aircraft (MIL-B-5087), and little radio-interference trouble due to arcing between poorly bonded members may be expected if these specifications are strictly adhered to. In fact, since there are many other reasons for bonding, some of which impose much more stringent requirements, strict adherence to the bonding specifications will usually insure the elimination of interference-producing arcs between members and parts covered by these specifications.

There are, however, certain cases, not fully covered by these specifications, which deserve special attention from the interference point of view. For example, arcing may occur between two metallic surfaces in the presence of a strong electric field, as near a transmitting antenna, if the bonding connection between them has opened up by corrosion. The resulting radio interference may well be the only indication of the fact that a poor bond exists. Another example is the arcing that may occur between the individual strands of copper-mesh screening at the cross-over points if good electrical contact is not made. It has been observed that the copper-mesh wall of a screened room can be the source of considerable radio interference when strong radio-frequency fields are present. Good bonding of the strands at the cross-over points eliminates this interference entirely. Therefore, it is concluded that the same construction techniques must be employed for copper-mesh screens used, for example, to shield ventilating louvres of electrical machines in aircraft.

3.2.3.5 SWITCHES

Any switching device causes transients during opening and closing, as explained in Paragraph 1.3.2.2. Therefore, in the design of all switches used in aircraft, the radio-interference problem must be considered from the outset.

As was pointed out in Paragraph 1.8.1.2, arcing occurs during the normal operation of a switch when used to interrupt the flow of current. In fact, the interruption of a current in a circuit may be said to consist of substituting a highly ionized and therefore conducting gaseous medium, i.e., an arc, for a part of the metallic circuit, and then subjecting this arc to strong de-ionizing influences. The arc is extinguished when the energy stored in the inductance of the circuit is dissipated and the voltage drops below the value required to maintain the arc. To prevent the arc, the current, instead of being interrupted, is channeled into another branch

containing a series capacitance and resistor. Thus the energy is partly stored in the capacitor and partly dissipated in the resistor, which also serves to damp out any oscillations that may occur as a result of the added capacitance.

The design of these R-C arc-suppression networks is treated in Paragraph 1.8.1.2. They must be used whenever switches or relays are used to interrupt currents large enough to cause radio interference. An alternative method is to completely shield and filter the unit, and when the currents to be interrupted are large, this may be the only effective method. But for small units, the use of an efficiently designed R-C network may make shielding and filtering unnecessary, and great savings in weight and space requirements may be effected.

3.2.3.6 FLUORESCENT LAMPS

Fluorescent lamps contain mercury vapor at low pressure which under electron excitation, obtained by applying a difference of potential across the lamps, emits invisible ultra violet radiation. This in turn excites visible luminescence in the internal phosphor coating. A fluorescent lamp ballast - a coil of insulated copper wire wound on an iron core - is placed in series with the lamp to limit the current to its rated value. Many lamps are also equipped with starters whose function is to complete a separate circuit so that a pre-heat current can flow through the cathode before the lamp operates.

Since basically the light source in a fluorescent lamp is an arc, considerable interference may be expected. This interference is both conducted away from the lamp through the power leads and also radiated directly from the lamp. The radiation takes place mostly at frequencies above 50 megacycles and is very difficult to control. Shielding with solid shielding material is obviously not feasible because there is no transparent conducting material. Shielding by means of copper-mesh has been attempted, but such shielding also reduces considerably the amount of light available from the lamp. The exact frequencies and intensities of direct radiation are quite unpredictable and depend greatly on the age, condition, and type of lamp. For this reason, the use of fluorescent lamps in the vicinity of any electronic equipment or sensitive wiring should be avoided entirely.

As far as the conducted interference is concerned, the capacitor normally connected across the switch contacts to aid starting also aids greatly in by-passing the interfering current components and keeping them out of the power line. If the interference reduction with this one capacitor is not sufficient, a power-line filter, designed according to the procedures explained in Paragraph 3.1.1.2, must be added. Thus it is seen that the use of fluorescent lamps is undesirable also from the point of view of conducted interference. However, the interference originating in fluorescent tubes is sometimes put to use as the signal source of signal generators useful in measuring receiver response in the ultra high frequency range. The signal generators presently in use are adequate for general testing but because of their limited power output they are inadequate for testing wide range receivers.

3.3 DESIGN CONSIDERATIONS APPLIED TO AIRCRAFT SYSTEMS

Present day military aircraft are almost completely controlled by electrical systems. Even hydraulic systems depend upon electrical control circuits for their

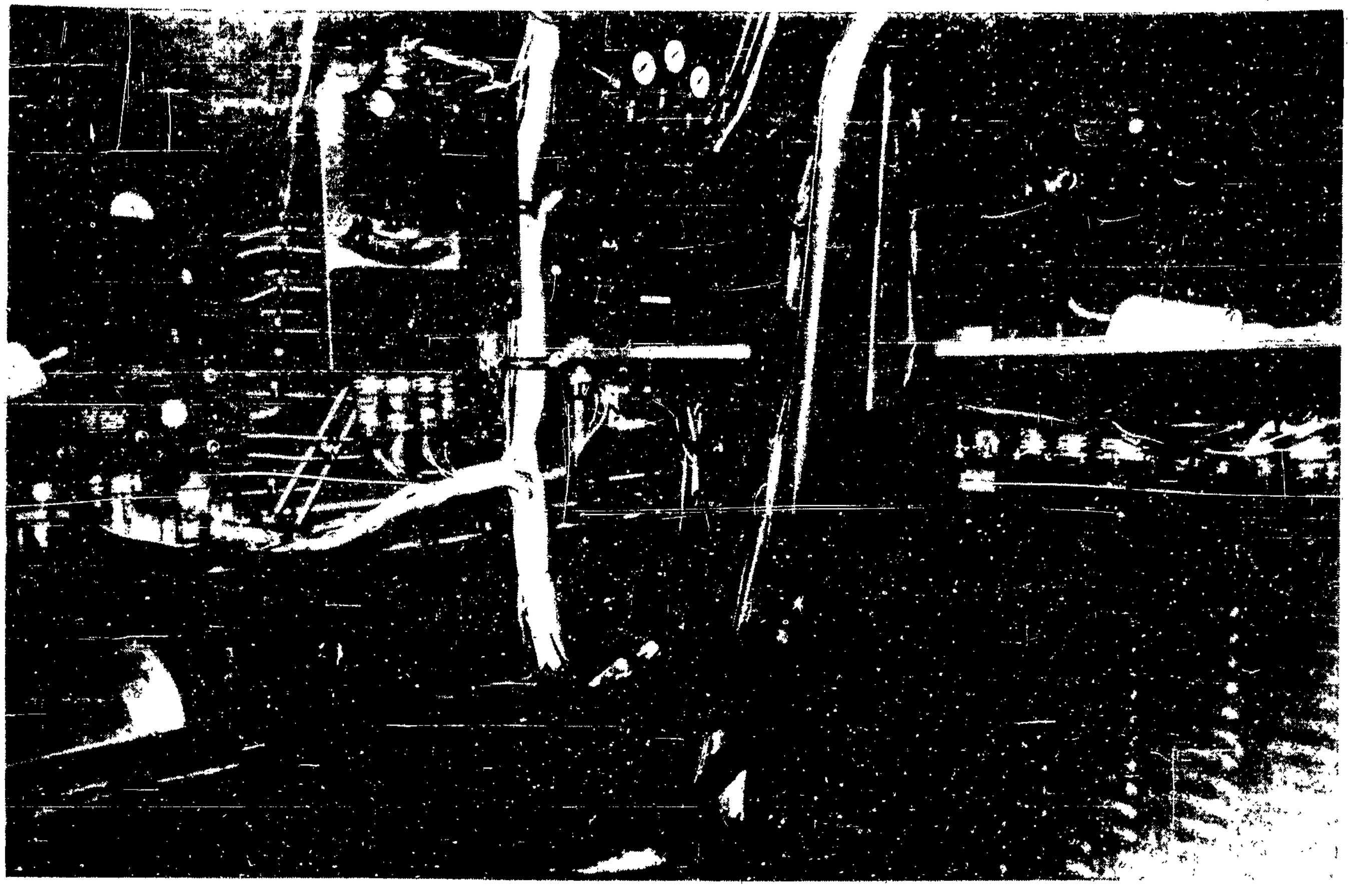


Fig. 3.3 - A Radar Operator's Position

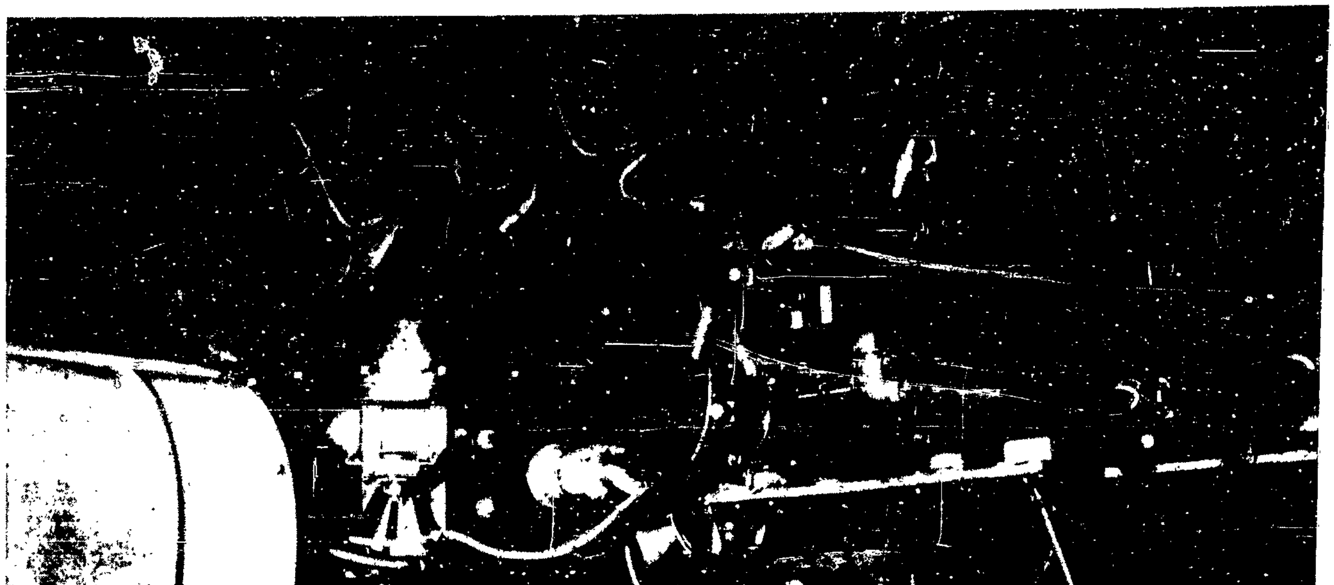


Fig. 3.3 - B Radar Compartment Installation

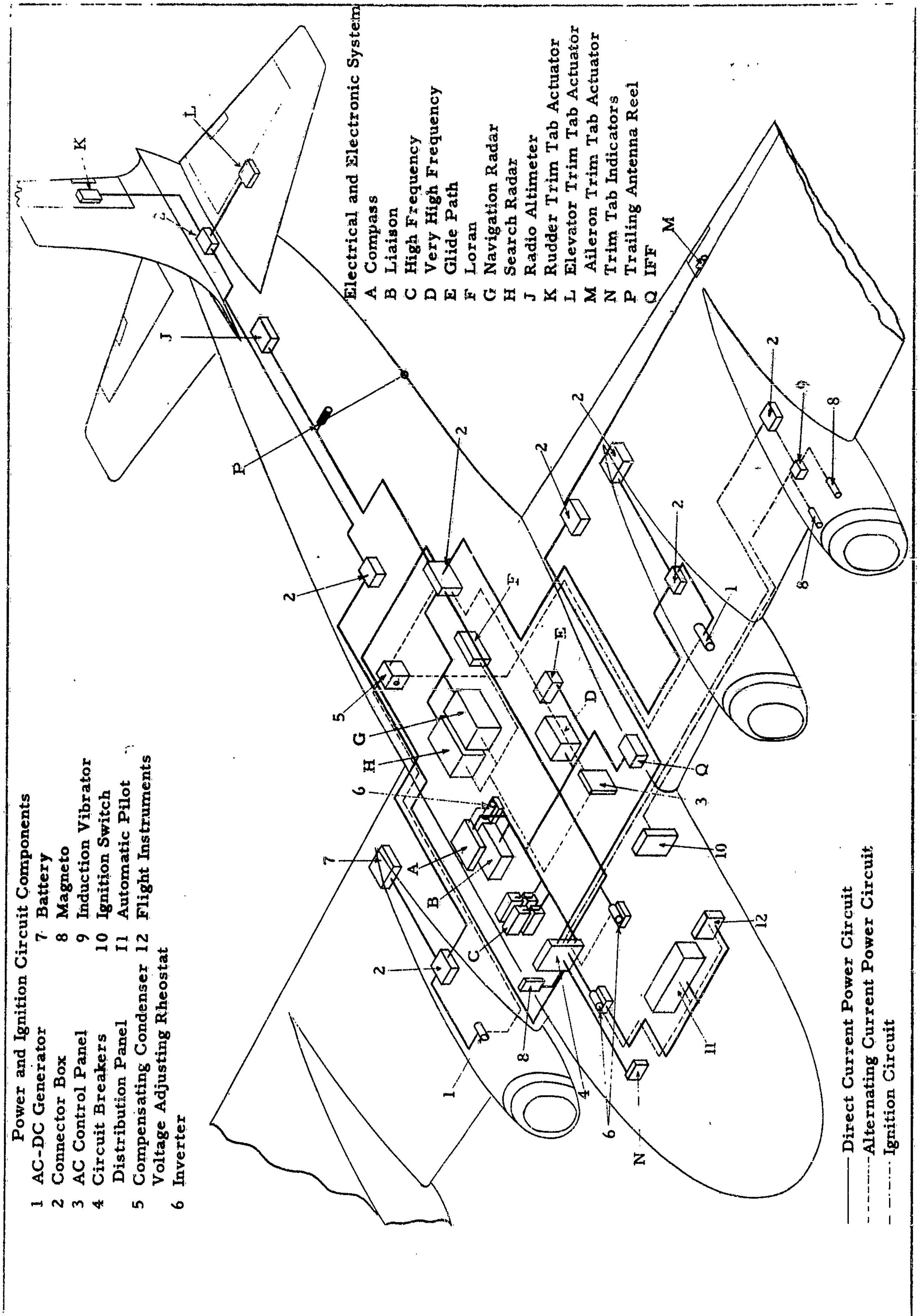


Fig. 3.3-C Typical Electrical and Electronic Power System

operation. The large increase in number and complexity of the electrical and electronic systems has greatly increased the radio interference problem. Close proximity of components, bundling of wiring into common cables, and high energy interference sources, have increased the number of paths over which interference signals may enter radio and radar receivers as well as the probability for such action. Radio interference levels that produced no adverse effect whatsoever in old model aircraft can no longer be tolerated in present day models. Since each crew member must operate one or more radio or radar sets, all indicators and controls involved must be located within the respective operator's reach. This requires bundling of power, indicator, and control wiring of several different systems in the same cables. Figure 3.3-A shows a radar operator's position in a typical aircraft installation. Here the close equipment mounting, parallel wiring, and wire bundling can be seen clearly.

A typical radar compartment installation is illustrated in Figure 3.3-B. The radar modulator, IFF unit, alternator, and inverter are located relatively close to one another and their interconnecting cables are bundled together. This offers the possibility for the high level radio interference emanating from the modulator unit to enter the IFF and AC power systems and gain access into many other electronic circuits throughout the aircraft. Also, the interconnecting cables passing through the compartment may pick up interference signals and conduct them into the various receivers. It should be stressed that electrical servo systems and control actuators may also introduce interfering signals into the electronic circuits.

A typical aircraft electrical and electronic system installation is represented in Figure 3.3-C. This shows the necessity for bundling and paralleling of circuit wiring. The operational necessity for mounting control boxes and indicators close together is also illustrated in this figure. When it is fully realized that each system must operate in conjunction with the many other systems in close proximity, then the magnitude of the system installation problem can be appreciated.

Design engineers of electrical and electronic systems must be thoroughly familiar with the installation problems and techniques to insure interference-free operation in the original lay-out. The system must be so designed that interference signals cannot enter or leave the system, due to conduction, radiation or inductive or capacitive coupling. Proper observance of good source suppression and systems design techniques will guarantee satisfactory functioning of all electrical and electronic systems regardless of their number or complexity.

3.3.1 ELECTRONIC SYSTEMS

Electronic systems are generally susceptible to interference although any one of them may be capable of generating interference which can affect some other electronic system. The arrangement and interconnecting wires of the component parts of each of these systems within the aircraft presents an array by which interference may couple into the system. An analysis from the radio interference point of view of each of these systems can be made without any prior knowledge of how interference may actually couple into or leak out of the system. Various components of each system are either susceptible to picking up radio interference or generators of interference or harmonics. If the assumption is made that the source suppression of "noisy" components cannot be absolutely perfect, additional design considerations such as routing of wires, shielding, shading, arrangement and location of equipment

are necessary to insure interference-free operation of the system and should be pointed out to the design engineer to increase his appreciation of the overall problem. Moreover consideration should be given to reducing the susceptibility of various parts of the system so as to tolerate the presence of interference fields without any adverse effects. This in no way advocates a "tailored" installation. It is merely intended to illustrate to any individual design engineer why and how he must broaden his viewpoint of the radio interference problem and where in the design of the general layout of a system attention must be given to certain special considerations. Many specific examples exist where interference entered or leaked out of a system as a result of poor design practice. The following paragraphs describe typical installations which serve to point out some of the general considerations. Each of these has peculiar characteristics as to power supply, antenna location, receiver-transmitter combination, etc., which deserve enumeration and illustration to create a picture of the nature of the physical situation under consideration.

3.3.1.1 RADIO RANGE AND HF SYSTEM

High frequency radio systems used in present day aircraft are designed to provide air-to-ground communication utilizing voice, code, and tone modulated CW transmission. Essentially, the system is a multi-channel radio receiving and transmitting equipment used for HF command and radio range purposes consisting of (1) three receivers with frequency ranges of 190 - 550 kc, 3.0 - 6.0 mc and 6.0 - 9.1 mc; (2) two transmitters with 5.3 - 7.0 mc and 4.0 - 5.3 mc frequency ranges, (3) remote control boxes, and (4) a modulator, all located in the cabin on the top radio shelf in the navigator's compartment; (5) an antenna relay mounted in an inverted position on the cabin overhead, and (6) two wire antennas running between three masts located on the top of the fuselage, (one for the range receiver and the other for the transmitter and the command receivers). The HF equipment provides transmission on two preset channels and reception through the frequency range of 3.0 - 9.1 mc. The low frequency receiver receives radio range signals in the 190 - 550 kc frequency range. In a typical installation this system serves the following stations: (1) pilot, (2) co-pilot, (3) radio operator, (4) radar-navigator, (5) two observers, and (6) tail compartment stations.

The receivers and transmitters are potential radio interference sources and are also susceptible to radio interference. Since the components of this system are widely separated there is a strong possibility that interference signals can enter or leave the system by radiation or inductive (or capacitive) coupling from the system wiring. Conductive access to and from the HF system is also provided by the interphone and power systems. The paths over which interference can enter or leave the system as shown in Figure 3.3.1.1 are: (1) antenna leads, (2) power leads, (3) mechanical remote control cables, (4) interphone connections to the earphone and microphone, and (5) penetration of case. There is also the possibility of intersystem interference through the antenna relay.

Interference entering the transmitter case would be radiated and manifest itself at receiving stations on the ground or in other aircraft, in the form of disturbing modulations producing "noise" in the respective audio output systems. Since this type of interference has not been encountered in actual experience, the problem will not be discussed further except to point out the possibility and to recognize that these paths of entry do exist. Some future installations could alter the situation sufficiently

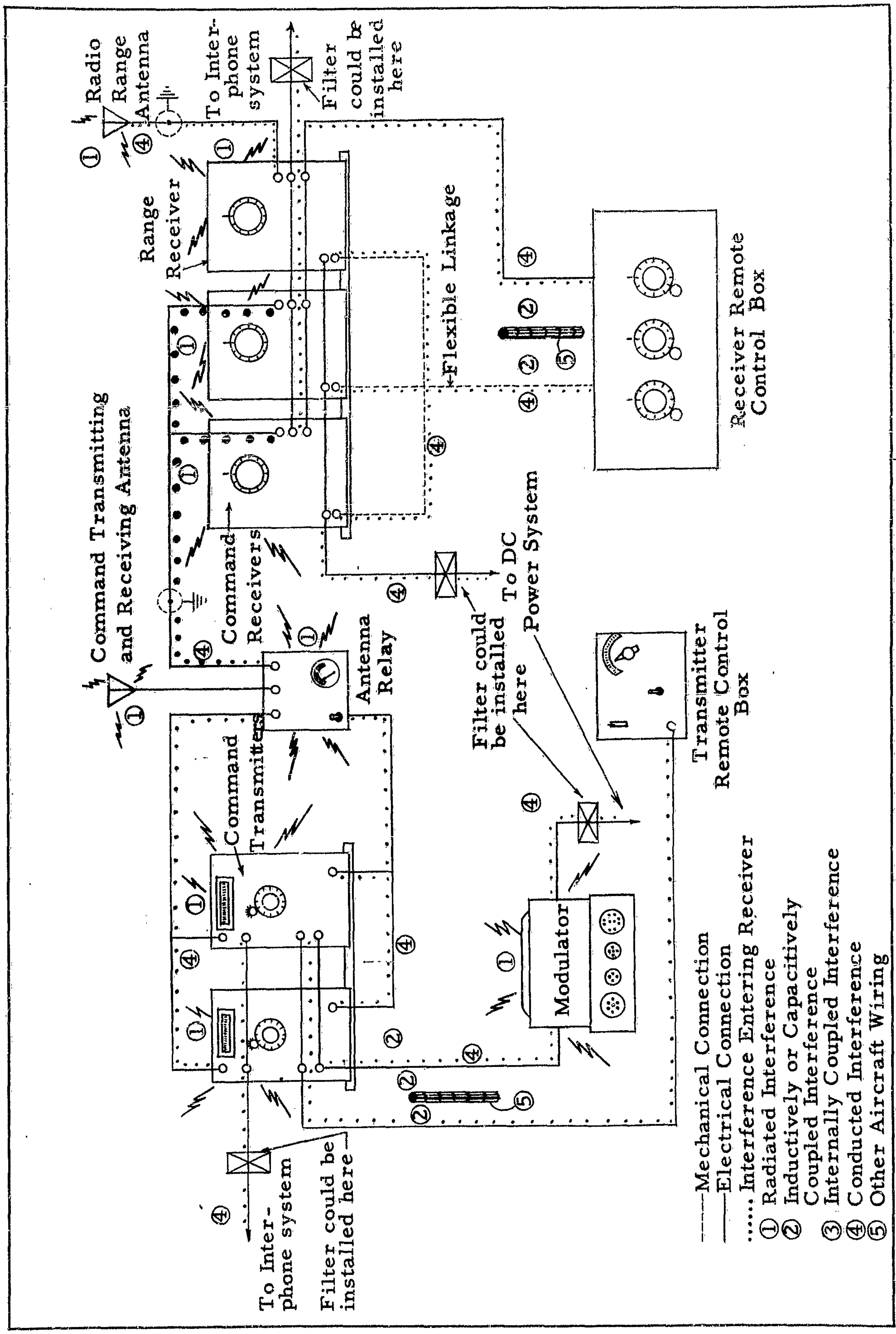


Fig. 3.3.1.1 Paths of Interference Signals in a Typical Radio Range and HF System

to produce malfunctioning in the transmitter or in ground stations if proper design techniques are not carefully considered during the functional design of the transmitter system.

The circuitry of the HF system should be arranged and components placed so that a minimum of shielding and filtering is required. Reference to Figure 3.3.1.1 shows that the HF system is directly connected to other electrical and electronic systems in the aircraft through connecting cables: (1) DC power supply to transmitter and receiver dynamotors, (2) receiver audio output to interphone system, (3) transmitter microphone connection to interphone systems. These lines should be filtered to prevent interference from entering or leaving the system by conduction (refer to Paragraph 1.8.2.3 and Appendix VII for details of filtering and filter design).

The long antennas, the relatively long antenna leads, and the long HF system wiring increase the susceptibility of the overall system to radiated interference as well as increase the interference-source potentialities of the system. This can be reduced by proper routing and shielding where necessary. In general, the antenna lead to the range receiver is adequately shielded and bonded to prevent interference difficulties over this path. However, if this antenna lead should be routed very close to a high energy interference source, additional shielding would be required. The command receiver leads are shielded between the receiver rack and the antenna relay so as to be relatively interference-free. The transmitting antenna lead between the transmitter rack and the antenna relay as well as the antenna lead from the antenna relay to the command antenna are not shielded and therefore introduce a serious interference problem.

The long HF system wiring extending the length of the fuselage in a typical installation provides a means of coupling interference signals into or out of the HF system. If these cables are bundled with other susceptible system wiring, or other probable interference - generating system wiring, they should be shielded, unless rerouting is practicable.

Satisfactory HF system operation in conjunction with the electrical and electronic systems in actual aircraft installations depends largely upon the care exercised by the designer in applying radio interference suppression techniques to the installations of other aircraft systems as well as the HF system itself. However, in any event, observation of design techniques outlined in this book together with specific attention to the particular problems discussed herein should result in a system free from objectionable radio interference.

3.3.1.2 VERY HIGH FREQUENCY RADIO RECEIVER SYSTEMS

Very high frequency radio systems used in present day aircraft are designed primarily to provide air-to-air or air-to-ground communication. The system is generally composed of a radio receiver, radio transmitter, power junction box, antenna, control units, and necessary interconnecting cords. In early models the transmitter, receiver, and power junction box were housed in separate cases while later more compact designs have incorporated all three components in one case.

Anv one of eight channels within the VHF range may be selected for operation.

Remote operation is provided by a remote control box and control cables. An audio output signal provided by the VHF receiver is available at any one of the interphone stations.

The installation of a representative VHF Radio Set in a typical aircraft has been selected as an example. The discussion to follow applies specifically to this particular layout. However, since all such systems are functionally similar, generality is still maintained.

One antenna is utilized to radiate or receive radio-frequency energy by the VHF system. This antenna is generally mounted on the top or underside of the fuselage, near the pilot's or navigator's compartment. The pilot's remote control box is mounted within reach of the pilot when seated at the flight controls. In a typical installation one interphone control box each, located within reach of each crew member, would be provided for the pilot, co-pilot, navigator, and radio operator. The receiver, transmitter, and power junction are mounted in the compartment behind the pilot. There are no components located in the tail or wing sections. Interconnecting cords are bundled with other electrical wiring passing through the compartment.

The three principal components, receiver, transmitter, and power junction box, are potential radio interference sources. Of these, only the transmitter and receiver are susceptible to radio interference. Even though the functioning of the control and power junction boxes may be unaffected by interference signals, they may serve as paths of entry for undesired signals.

As shown in Figure 3.3.1.2-A, interfering signals may enter the transmitter case by way of the (1) antenna terminal, (2) antenna connection from the receiver, and (3) power and control cables.

Interfering signals may enter the receiver case over any of the following paths also shown in the figure. These are: (1) antenna terminal, (2) power and control cable, or (3) through penetration of the case. A detailed discussion of receiver design techniques for interference-free operation is given in Paragraph 3.4.

Interference entering the transmitter case may be radiated and manifest itself at receiving stations on the ground or in other aircraft, in the form of disturbing modulations producing "noise" in the respective audio output systems. Since this has not been particularly disturbing, the problem will not be discussed further except to point out the possibility and to recognize that these paths of entry do exist. Some future installation could alter the situation sufficiently to produce malfunctioning in the transmitter or in ground stations if proper design techniques are not carefully considered during the functional design of the transmitter system.

Transmitter case radiation and harmonic generation have caused considerable difficulty in aircraft installations. The design of the transmitter housing in the typical case selected, did not include screening of the ventilating louvres. The inspection

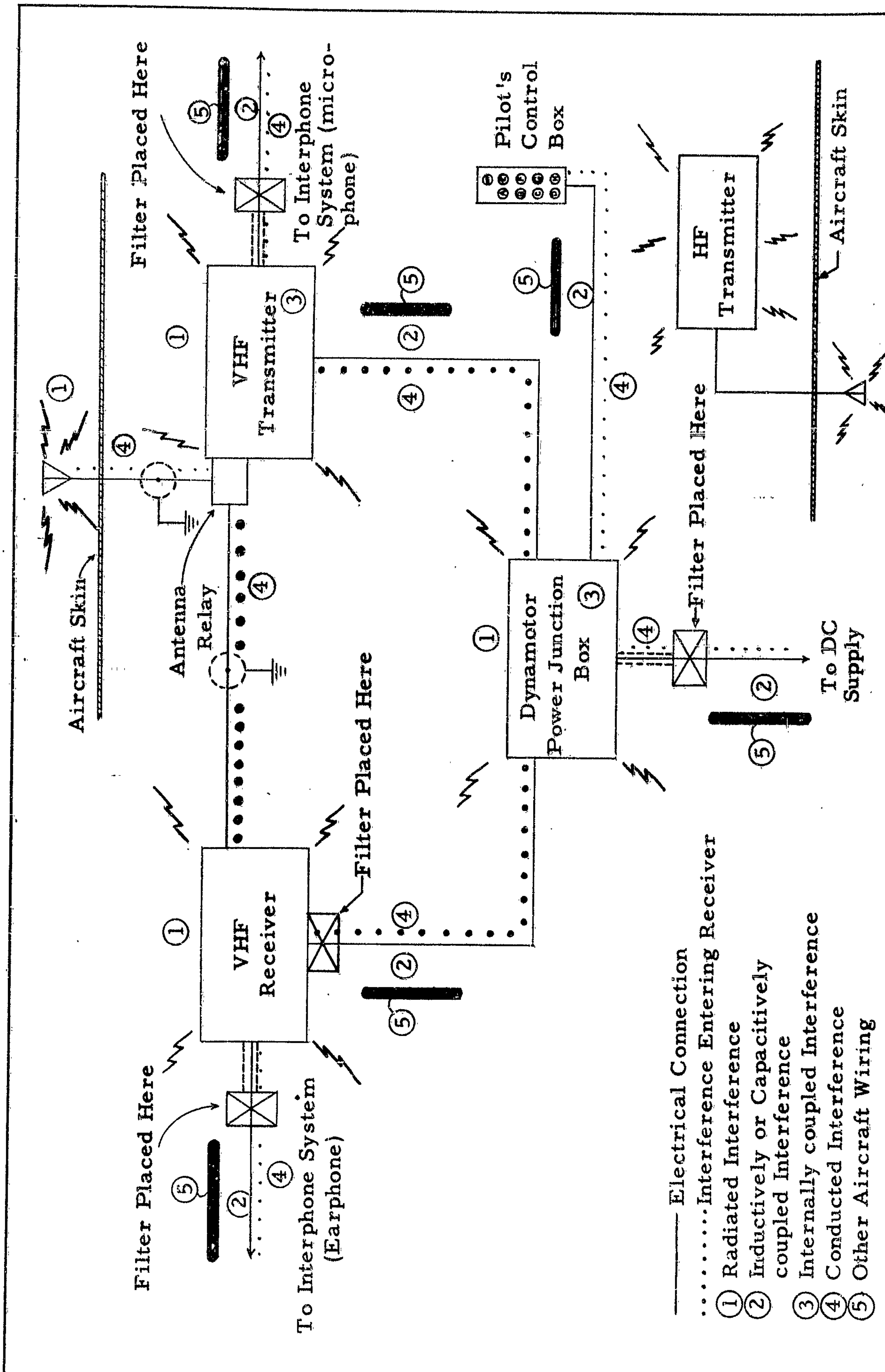


Fig. 3.3.1.2-A Paths of Entry of Interference Signals in a Typical VHF Communication System

plate on the base of the chassis made poor contact with the case and the electron tubes with glass envelopes were radiating appreciable energy. It was necessary to incorporate the following modifications in order to reduce the interference generated by the transmitter to within tolerable limits: (1) shield and bond first tripler and final stages, (2) rivet a metal screen to the inside of the cover to reduce radiation through the ventilating louvres, (3) bond the inspection plate on the base of the chassis with a copper gasket, (4) install a bonding strap around each shock-mount, and (5) replace troublesome glass tubes with metal tubes. These difficulties would not have appeared in the original installation if proper component design had been observed. A detailed discussion of interference-free transmitter design is given in Paragraph 3.2.2.2.

The VHF receiver has caused interference with other electronic systems in the aircraft. Automatic selection of the various channels by an electrically operated, motor-driven, channel-selecting mechanism occurs when any one channel push button is depressed on any remote control box. This band-change motor caused excessive interference over a wide band of frequencies. The interfering signals were present on all interconnecting cables and also appeared at the supply terminals. Since the interference was present only during the "warm-up" period and during the band-changing period, no corrective measures were attempted in the typical case selected for discussion. However, it should be emphasized that this type of interference need not appear in a system, even for short time intervals, if proper component design techniques were applied during the original design to suppress the interference at the source. A detailed discussion of interference-free design techniques for small motors and receivers is given in Paragraphs 3.2.1.1 and 3.4.

Considerable interference was generated within the VHF system by the dynamotor power unit in the power junction box. Filters were installed in the power junction box to suppress the dynamotor commutator interference at the source. It was also discovered that the high voltage leads were exposed to radiation from the dynamotor after leaving the filter. Rerouting these leads reduced the interference to a permissible level.

In the original installation, components should be placed, and the circuitry arranged so that a minimum of filter components are necessary. Reference to Figure 3.3.1.2-A shows that the VHF system is directly connected to other electrical systems in the aircraft through connecting cables: (1) receiver audio output to interphone system, (2) transmitter microphone connection to interphone system, (3) DC power supply to power junction box. These lines can be filtered to prevent interference from entering or leaving the system by conduction (refer to Paragraph 1.8.2.3 and Appendix VII for details of filtering and filter design).

An interesting interference problem was encountered with this system which illustrates some poor design techniques. Interference was noted in this VHF receiver whenever the HF transmitter of another system was operated at 12 megacycles, the IF frequency of this receiver. Disconnecting the receiver antenna did not eliminate the interference. The test described below proved the interference

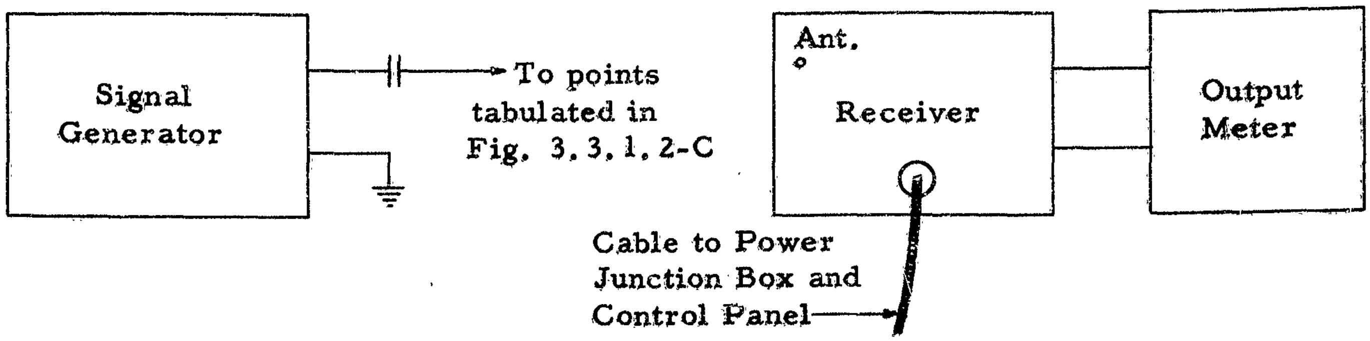


Fig. 3.3.1.2-B Test for Source of Interference in Receiver

Lead Wire Function	Signal Generator Output In volts	Receiver Output In volts
Antenna Post	0.32	1.55
Channel Selector A	0.26	8.5
Channel Selector B	0.23	1.7 *
Channel Selector C	0.26	10.2
Channel Selector D	0.27	2.0 *
Channel Selector E	0.26	8.0
Channel Selector F	0.26	8.0
Channel Selector G	0.27	10.3
Channel Selector H	0.27	10.4
+200v	0.26	10.4
Audio Low	0.26	7.0
Time Delay	0.26	10.4
Ground	0.26	1.1 *
+28v	0.27	10.6

* Denotes wires which were not filtered. All other wires in this cable were filtered to prevent the 12 mc signal from entering the receiver.

Fig. 3.3.1.2-C Susceptibility Tests of VHF Receiver

was entering the receiver via the power and control leads. The signal generator was tuned to 12 megacycles and the modulation was 30 percent at 1000 cycles per second. The signal was applied to the antenna post and the individual wires in the cable shown in Figure 3.3.1.2-B with the results tabulated in Figure 3.3.1.2-C. The background interference level at the output was 0.65 volts with the antenna disconnected and no signal applied.

In this particular installation, the leads were filtered to eliminate the interference. Two faulty design practices created the need for ten filters; however, if these had not existed, the added weight, cost, etc., would not have been required. They are:

(a) The HF transmitter antenna lead-in was unshielded and rather long. The 12 mc signal radiated by the lead-in was picked up on the control wires of the VHF receiver and interfered with its operation. The antenna lead-in had to remain unshielded because the transmitter output circuit was not designed for the load presented by a shielded lead-in.

(b) The VHF receiver circuits were interference susceptible due to coupling between the IF wiring and the control and power wiring. It should be noted that the IF rejection of the antenna input circuit was far better than the equivalent rejection for the susceptible wires in the cable.

Overall design considerations for the system could be improved by reducing the number of separate components with the consequent elimination of external wiring. Some of the later model VHF systems have incorporated the receiver, transmitter and power unit in one case. This compact construction is advantageous from the standpoint of shielding, wiring, and overall interference-free design.

The VHF antenna system is provided with a feed-through insulator, coaxial cable to transmitter, a switching relay in the transmitter, and a coaxial cable to the receiver. This type of antenna design is in agreement with sound interference-suppression techniques and actual installation experience has proved this arrangement satisfactory.

This system serves as an excellent example to point out the necessity for thoroughness in considering the interference problems in the design of any system. Poor transmitter and dynamotor component design as well as poor systems shielding and filtering have produced a system that is a relatively strong radiator of interference signals. If good interference-suppression design techniques had been employed in the overall system as well as to the components of the system, the need for the system fixes described above would have been eliminated and a satisfactory operation would have resulted in the original installation.

3.3.1.3 SEARCH RADAR SYSTEMS

Airborne search radar systems are designed to present a visual representation

of a portion of the earth's surface or objects on a radar scope to supply accurate navigation and bombing information independent of weather and visibility conditions. This system is generally composed of (1) an antenna assembly, (2) transmitter-receiver unit, (3) modulator, (4) indicators, (5) synchronizer, (6) rectifier power unit, (7) junction boxes, (8) control units, (9) gyroscope, (10) directional coupler, (11) blowers, (12) servo amplifier, (13) camera attachment, and, (14) various interconnecting cables. The antenna is operated by a servo motor so as to scan a portion of the area under and around the aircraft. Circuits, controls, etc., are arranged to radiate radio frequency pulses, synchronized with the indicator sweep circuits, into the region searched and to receive the returning reflected signals or echoes. These received echoes vary in strength because of the different reflecting properties of the objects in the area scanned. Each received signal is converted by the equipment into a light spot on a cathode-ray indicator. Since the intensity of the light spots are dependent upon the amplitude of the received signal, a light and dark map-like pattern appears on the radar scope. There is no audio output signal from the system. The antenna assembly is generally mounted on the underside or in the nose of the aircraft. All other components are mounted in the compartment behind the pilot where they will be accessible to the navigator or radar operator. Due to the large number of components, considerable interconnecting wiring is required and a large portion of the system wiring is bundled with other electrical and electronic system cables that pass through the compartment.

The installation of a representative model search radar set in a typical aircraft has been selected as an example for discussion of design techniques for interference-free operation. In this system, the receiver output is a visual pattern displayed on the radar indicator scope, and any unwanted signals which find their way into the radar receiver case must be capable of disturbing the indicator pattern in order to constitute an interference problem. In Paragraph 1.7.1, the nuisance value of interfering signals in various types of receivers is discussed in detail.

The components in a typical search radar installation that should be given special attention as potential interference generators are: (1) antenna assembly, (2) motor driven fans, (3) receiver-transmitter unit, (4) indicators, (5) synchronizer, (6) modulator, (7) power unit, (8) servo amplifier, and (9) the gyroscope unit. Of these units, those that can also be classed as "receivers" under the extended definition of receiver given in Section I are items (3), (4), (5), and (8). Even though the functioning of the other components of the system is unaffected by interference signals, they may serve as paths over which interference may enter or leave the system.

Figure 3.3.1.3-A illustrates the paths of interference signals in a typical search radar system and shows the most vulnerable points for interference signals to enter the system. Interfering signals have gained access to the various susceptible components over the paths shown and have eventually exercised sufficient influence on the indicator pattern to constitute an interference problem.

The location of the antenna assembly of a radar search system is particularly important. When the general design of the aircraft requires mounting the antenna in a location where the radar receiver-transmitter field to the sides or to the rear is blocked by reflecting surfaces, standing waves of high amplitude will generally be established in the antenna system. This results in faulty operation of the radar

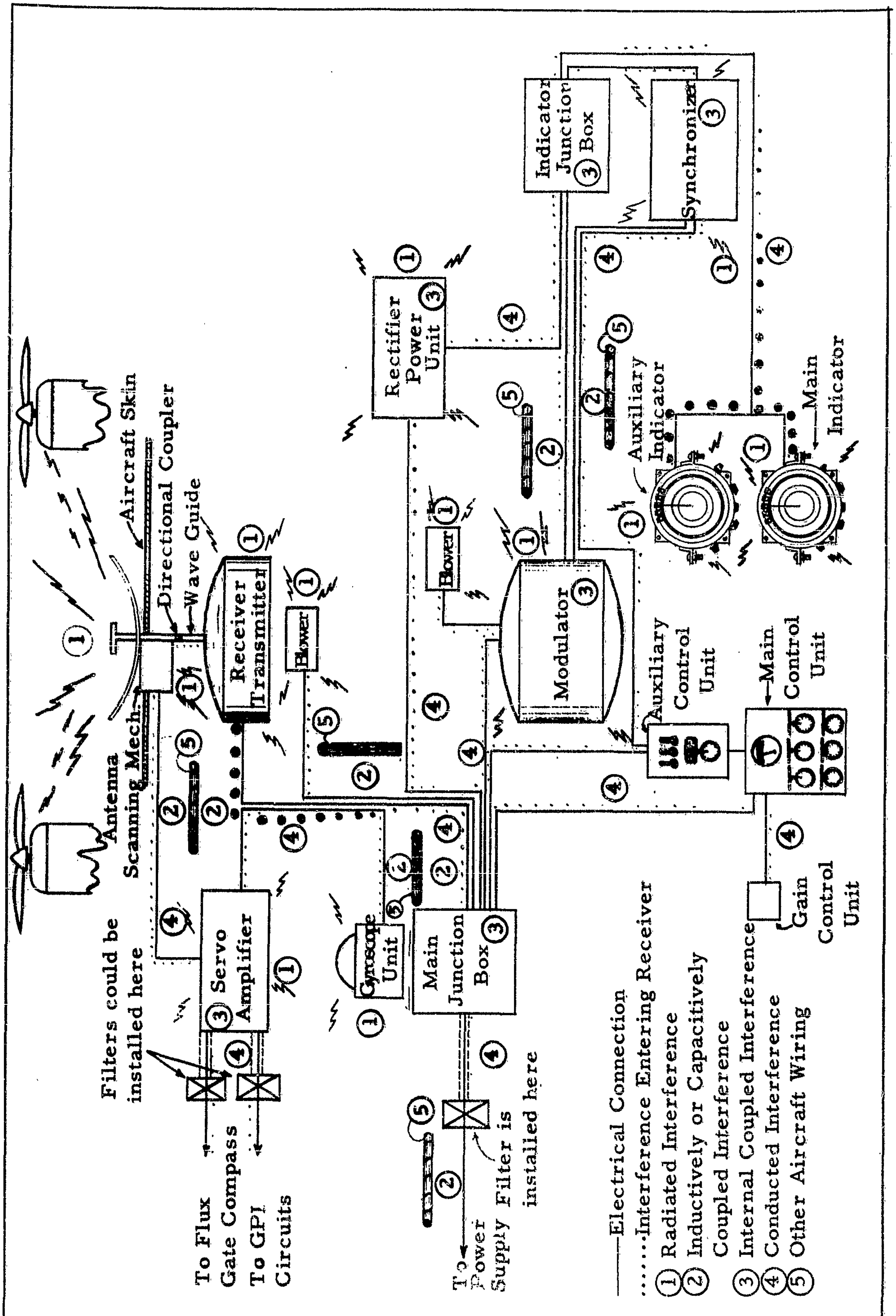


Fig. 3.3.1.3-A Paths of Interference Signals in a Typical Search Radar Set

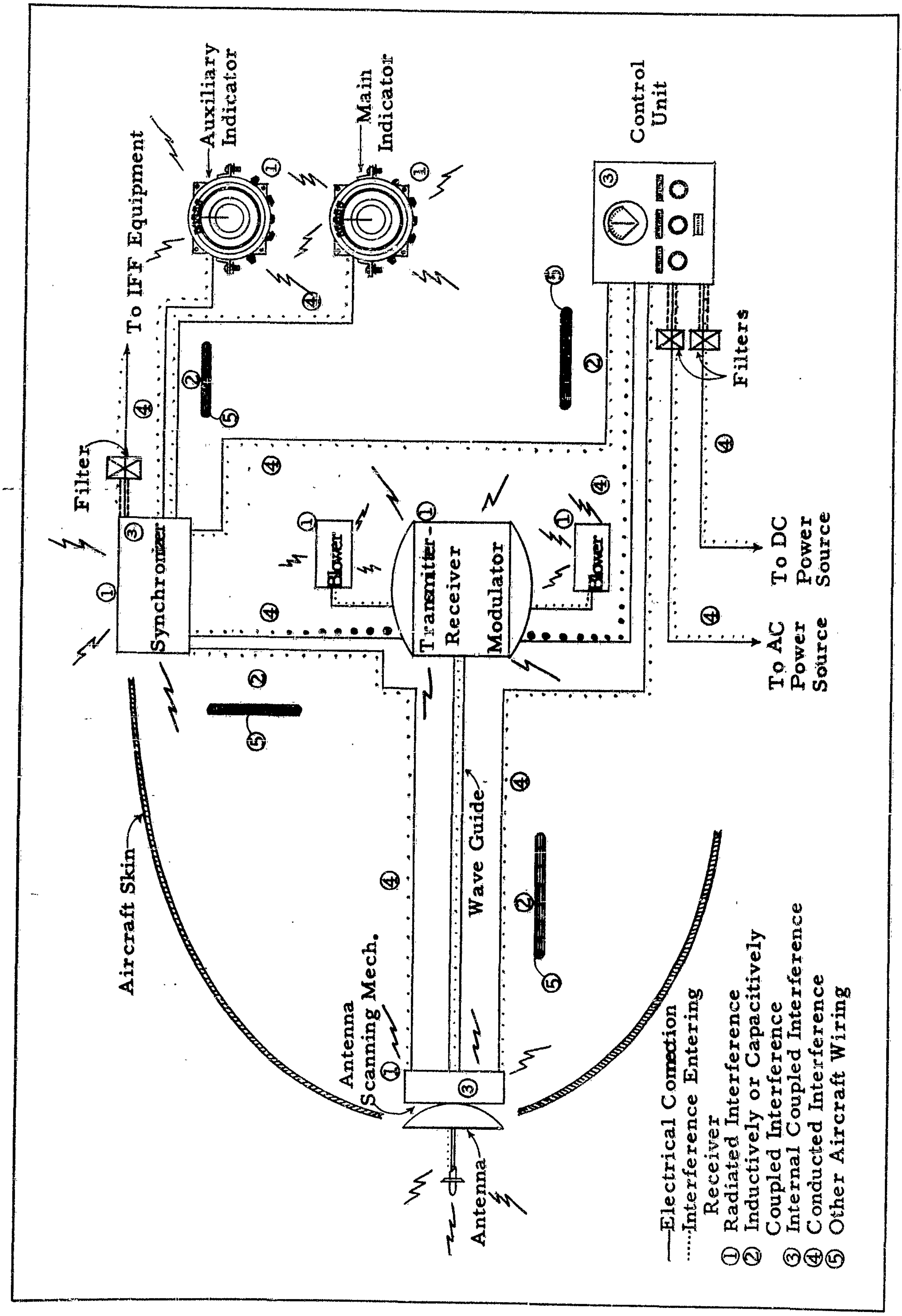


Fig. 3.3.1.3-B Paths of Interference Signals in an Improved Version of a Typical Search Radar Set

set. Considerable reduction of this type of radio interference can be obtained by coating the reflecting surfaces involved with material to absorb as much radio frequency energy as possible. Such measures are essentially the only means available for eliminating the disturbance when relocation of the antenna assembly is not feasible.

Interference in this typical installation entered the radar system also over the power cables. This disturbance was suppressed by inserting a filter at a convenient position between aircraft power supplies and the radar system junction box. Some difficulty was also encountered when the cable from the control unit to the synchronizer and the cable from the synchronizer to the indicator were mounted too close to the power cables of either the radar system or other aircraft wiring. This was corrected by separating the power cables and the radar system indicator or synchronizer cables by at least 18 inches when installing the system wiring.

Radio interference tests have demonstrated that search radar systems introduce a severe interference problem in other aircraft systems. In a typical installation the search radar pulse modulator circuits caused interference in the liaison and radio compass systems. Extended filtering and shielding was ineffective. Probing of the radar cabling indicated a high interference level over extensive lengths of various cables. This suggested that external filtering would be very difficult and that source suppression within the modulator together with the rerouting of certain portions of the cabling would be the most efficient solution. For a discussion of modulators refer to Paragraph 3. 2. 2. 3.

Radio interference from the computer appeared in the audio output of the radio compass. Shielding of the radio compass sense antenna was required to prevent interfering signals from entering the radio compass receiver.

In the typical case under discussion, intermittent interference signals caused by the determination-switch operation and tracking-control operation appeared in other electronic systems on the sircraft. Attempts to filter all radar power cabling at the power connections proved useless. As shown in Figure 3. 3. 1. 3-A, interfering signals from remote units must travel through long unshielded cables to reach the filter. This permits coupling of interference into other aircraft wiring as well as coupling into the search radar circuits. In addition, all power wiring had wire return circuits from the power unit to the main filter, and filters were provided for these wires just prior to grounding. This indiscriminate use of filters served no useful purpose, unnecessarily increased the length of the system's wiring, and required the use of unduly large connectors to handle the extra pins. Here again, source suppression would have more efficiently guaranteed an interference-free system in the original installation.

Reference to Figure 3.3.1.3-A shows that the Search Radar System is directly connected to other electrical systems through the AC power supply cable into the main junction box. A filter is provided in this line to prevent interference from entering or leaving the system by conduction. However, the greatest interference problems in this typical search radar system were caused by coupling into other aircraft wiring. This coupling was traced to the facts that (1) the modulator unit was not adaquately bonded to the airplane structure, (2) surplus pulse cable was coiled and clamped next to aircraft wiring, (3) the AC lead from the modulator to the external blower was shielded, but "grounds" on the shields were too long, (these leads were "hot" from

the modulator pulse and it was necessary to shorten leads to one inch), and (4) the coaxial antenna cable for the Loran receiver passed through the interference field around the modulator, (although grounded at both ends, the cable provided an efficient path for conducting the interference to the liaison receiver because both receivers used the same antenna), and (5) coupling also occurred inside the liaison transmitter since the external power wires leading to the transmitter were exposed to the interference field and the interconnected internal wiring was in close proximity to the receiver antenna-grounding lead. Appreciable improvement in future designs and installations can be obtained by the application of the following procedures to obtain interference-free operation:

(a) Ground all four pulse cable shields at the connectors. (It was found that in some cases only one of the shields was grounded because of greater ease of assembly, a practice which resulted in the loss of much of the shielding effectiveness.)

(b) Install a suitable filter in series with the AC lead from modulator to external blowers.

(c) Install the modulator and receiver transmitter as far as possible from other radio equipment, especially the liaison and radio compass receivers. These receivers should be in a separate compartment if possible.

(d) The pulse and high voltage cables should not be bundled with, run parallel to, or placed less than one foot from other radio and aircraft wiring. At least eighteen inch separation should be maintained for cable lengths over twenty feet.

(e) Keep all wiring associated with the radar set well isolated from other aircraft wiring.

(f) Locate modulator and receiver-transmitter units so that pulse cable lengths are held to a minimum.

(g) Since the pulse cables are prefabricated in fixed lengths, extra cable is sometimes coiled to take up surplus. In case this is necessary, the coil should be placed in the bomb-bay or other isolated compartments and adequately spaced from other wiring.

(h) All radar components should be properly bonded to the aircraft structure by application of the techniques discussed in Paragraph 3.1.3. The modulator and receiver-transmitter units should have at least two such bonds of shortest practicable length.

(i) The AC lead from modulator to external blowers should be filtered with a portion of the lead between modulator and the filter shielded and grounded with short leads.

(j) In case interference is encountered due to penetration or leakage of the pulse cable shield, the interference may be reduced by grounding the shield at intervals of approximately 5 feet.

Some of the later models of search radar equipment have incorporated improved design techniques resulting in a considerable reduction in the interference trouble described above. Much of the difficulty caused by long cable lengths and modulator radiation have been eliminated by placing the modulator, transmitter and receiver in one metal case. Fewer component parts, reduction of interconnecting wiring and better relative location of circuit elements have also improved the overall design from an interference standpoint. However, even with these improvements, considerable trouble is still caused in other electronic systems by search radar equipment. The high energy pulse output of the radar constitutes a prolific source of interference. In an improved version of a typical search radar system, shown in Figure 3.3.1.3-B, the paths for interfering signals to enter or leave the radar system are indicated. Some of the interference problems encountered in the improved version and the modifications made in an effort to attenuate interference are discussed below:

(a) Excessive levels of interference were present in the high voltage, modulator pulse circuits which coupled interference to all interconnecting wiring and coaxial cable within the receiver-transmitter. This interference was decreased by the addition of filters and by-pass condensers on the interconnecting wiring, and by the substitution of double-shielded coaxial cable for the single-shielded cable. The leads to the external blowers were shielded to confine the interference to the receiver-transmitter unit. The grounding of the video and AN connectors to the junction box were improved by removing the paint from the connectors. Refer to Paragraph 3.1.3.1 for a detailed discussion of direct bonding. Furthermore, the mating surface between the receiver-transmitter lid and case was cleaned in order to obtain a direct metal-to-metal contact.

(b) Interference was generated within the synchronizer unit by the transients in the sweep and intensity-gate signals for the indicators as well as in the steep pulse for triggering the modulator. The level of interference was decreased by shielding the lines within the unit and the 120-volt leads to the antenna assembly and the two indicators.

(c) A high level of video interference in the indicators was caused by thermal agitation and by the shot effect in the receivers and was coupled by induction into the interconnecting cables from the two B+ leads between the indicators and the synchronizer. This interference was attenuated by the insertion of an R-C decoupling circuit, similar to the one discussed in Paragraph 3.4.3.1, in the B+ leads. Interference emanating from the case of the indicator was attenuated by shielding and grounding the gain control potentiometer and grounding the case of the intensity-control potentiometer. Further attenuation of interference was obtained by grounding the indicator through a shock-mounted bonding jumper. Refer to Appendix XV for a detailed discussion of shock-mounted bonding jumpers.

(d) Sparking at the commutators of the direct-current motors and the make and break action at the sector-scan switch points were the sources of interference within the antenna assembly. The motors were shielded to decrease the level of radiated energy and a line filter was used to attenuate conducted interference. Capacitor spark suppressors and filters were employed to attenuate interference generated by the sectoring switches.

From the above discussion it can be seen that this radar system continues to be a high level interference source in spite of the improvements. Although the corrective measures taken reduced the level of interference considerable, the radar set failed to meet the requirements of Specification No. 16E5(Aer) and MIL-I-6181. In order to meet the specifications the system should be redesigned with the following interference-free design features incorporated:

(a) The modulator unit should be totally enclosed within a mesh shield whose joints have been welded or soldered. Refer to Paragraph 3.1.2.1 for a detailed discussion of a mesh shield. To provide a low, radio-frequency impedance path to ground, the shield should be clamped or screwed to the chassis. Furthermore, all power and control leads should be filtered by the use of a feed-through capacitor at the point of entrance. The superior attenuation characteristics of feed-through capacitors are discussed in Paragraph 3.1.1.5. This method of shielding and filtering will confine all interference generated by the modulator pulse circuits to the modulator unit. All other circuits in the receiver-transmitter unit should be relatively interference-free.

(b) The use of inherently interference-free components in the radar system should be stressed. Alternating current induction motors, rather than fractional horse-power DC motors, should be used whenever possible. A detailed discussion of commutation, the worst offender of all sources of radio interference in rotating machines, is given in Paragraph 3.2.1.1.2. When a fractional horsepower DC motor must be used, it should be completely shielded and equipped with a line filter as explained in Paragraphs 3.2.1.1.3 and 3.2.1.1.4. Vacuum tube switching circuits are preferred to vibrating contacts such as relays or vibrators because these devices generate high levels of interference. If the use of relays or contacts cannot be avoided, the unit should be enclosed in a metal shield and all connections should be filtered.

3.3.1.4 INTERPHONE SYSTEM

Combat interphone systems are designed to provide interphone communication between the various stations of a multiplace aircraft. Switching facilities are provided by means of jack boxes at each interphone station to enable the crewman to exert partial control over the radio system required for the discharge of his duties. Examples of radio systems which can be partially controlled at the various stations in addition to the interphone system are: VHF, Liaison, Command, and Radio Compass.

The major components in a typical interphone system include an interphone amplifier, dynamotor, jack boxes, headphones, microphones, microphone switches and control panels. The quantity and type of components used in a specific installation is dependent upon the tactical needs of the aircraft.

Paths of entry of interference are shown in Figure 3.3.1.4-A. Interference problems arising from typical aircraft installations are given below.

The dynamotor mounted on the chassis of the interphone amplifier causes

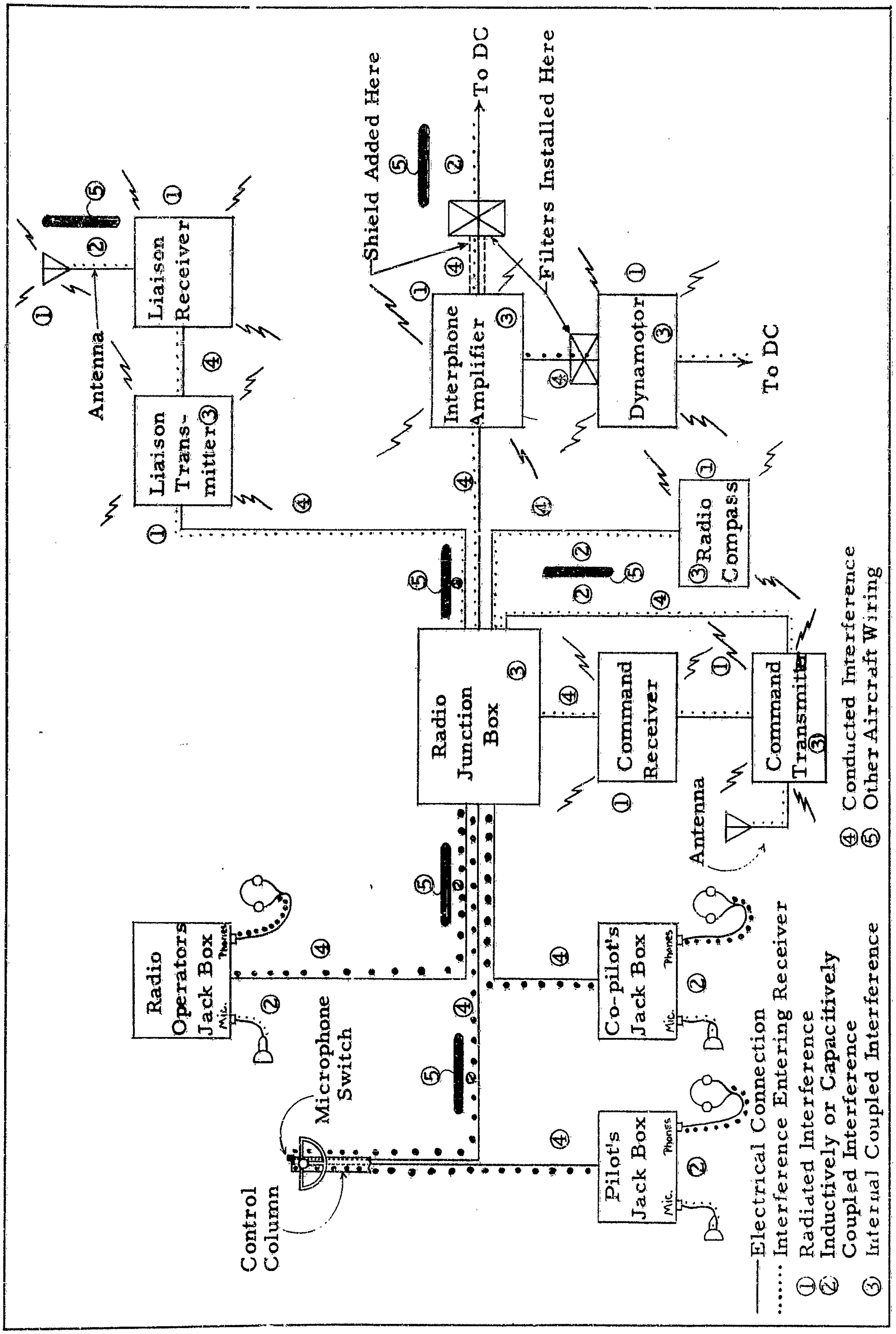


Fig. 3.3.1.4-A Paths of Interference Signals in a Typical Interphone System

interference in the output of many radio-frequency amplifiers incorporated in the various electronic systems due to the conduction of interference through the common power supply. This interference was satisfactorily attenuated by the use of a properly designed radio-frequency filter. Audio-frequency interference which was extremely apparent in the output of the interphone amplifier was decreased to a sufficiently low level by bonding the dynamotor to ground. Refer to Paragraph 3.1.3 for a discussion on bonding. Dynamotors, small units designed to convert direct-current power from one voltage magnitude to another, are prolific sources of both audio and radio-frequency interference primarily because they contain two commutators. Refer to Paragraph 3.2.1.1.2 for a discussion on commutation.

Interference was found to enter the interphone amplifier by conduction through the direct-current bus where ripple voltages as high as 4.4 volts at a frequency of 4000 cycles per second were measured. This interference, originating in the surface-control, hydraulic boost pump motors, was effectively attenuated by the insertion of a filter illustrated schematically in Figure 3.3.1.4-B.

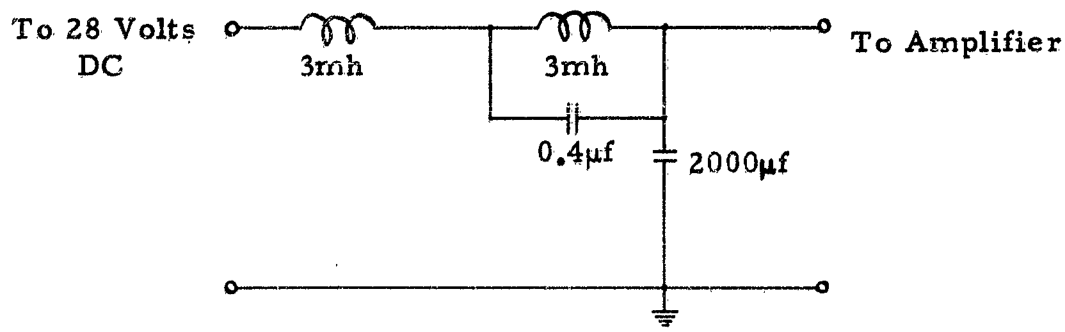


Fig. 3.3.1.4-B Line Filter for Surface Control Motors

The preceding problems were a result of interference being conducted into or out of the system. The following examples illustrate problems caused by interference coupling into or out of the system.

(a) Interference, coupled by induction from the power wiring to the interphone amplifier, was prevented by maintaining a minimum spacing of 50 inches between the power wiring and the amplifier.

(b) Inductive and capacitive coupling of interference from the power wiring into the interphone wiring of a specific aircraft installation was eliminated by replacing the original, single-wire interphone system with a two-wire system. The two wires were twisted, enclosed in a metal shield, and a minimum distance of 12 inches maintained between the shield and the power wiring. In general, the results of interference tests indicated that for greater interference-free operation all single-wire systems should be replaced by two-wire systems. Variations of the two-wire system have been satisfactorily used depending upon the function of the aircraft. For example, satisfactory operation of an interphone system in cargo planes, which are relatively free of interference-generating devices, has been accomplished by the use of two unshielded wires. In contrast, satisfactory

operation of the same system installed in bombing aircraft could only be accomplished by the use of two wires enclosed within a shield. The wiring systems employed in aircraft installations in order of their effectiveness in suppressing interference are:

(1) Two twisted wires in a common shield.

(2) Two parallel wires in a common shield.

(3) One shielded wire with a common return path, or two unshielded wires.

(4) One unshielded wire using the structure as a return path.

(c) Interference was present in the interphone system when the pilot's microphone switch was opened but disappeared when it was closed. This was caused by the coupling of interference by induction to the microphone switch wires which were run through the control column remote from the microphone as shown in Figure 3.3.1.4-C. Interference was attenuated to a certain degree by grounding the column, but much greater improvement was obtained by using a relay in the pilot's microphone wiring and by running relay control leads rather than the microphone switch wiring in the control column.

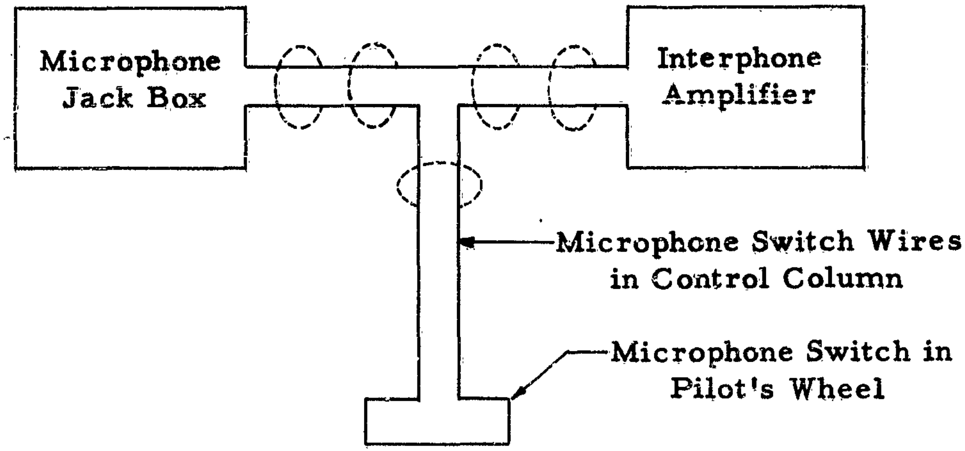


Fig. 3.3.1.4-C Routing of Microphone Switch Wires

(d) The dynamotor mounted on the interphone amplifier chassis was the source of interference present in the liaison receiver in a certain installation. Interference was coupled from the power leads of the dynamotor into adjacent aircraft wiring and eventually coupled into the unshielded antenna leads of the liaison receiver. This interference was attenuated to a satisfactorily low level by the insertion of a filter in the power leads of the dynamotor.

(e) In another typical installation, the liaison transmitter was mounted in close proximity to the interphone amplifier. Interference was inductively coupled

to the audio input transformer of the amplifier from the armature windings of the dynamotor employed as the power source of the transmitter. The level of interference was decreased sufficiently by enclosing the transformer within a metallic shield.

(f) The close proximity of the microphone and headset leads resulted in capacitive coupling between the input and output circuits of the interphone amplifier and caused audio oscillations in the interphone system. In order to prevent any oscillation from occurring, the amplifier components were connected so that the output and input circuits were as far out of phase as possible throughout the audio range of 100 to 20,000 cycles per second. This phasing was accomplished by connecting the secondary of the output transformer so that the signal voltage across the output lead and ground is out of phase with the signal voltage impressed across the input lead and ground.

The most recent design of an interphone system features an individual amplifier at each interphone station rather than a central amplifier. As a result, the audio signal is amplified at the source and conducted at higher levels and requires a minimum of amplification at the other interphone stations. This prevents the low "noise" levels in the lines from being amplified to an objectionable degree. Furthermore, the high level lines (high signal-to-interference ratio) permit the use of dynamic microphones which increases intelligibility by 65 to 95 percent. Boom, mask, and hand-held microphones are examples of the interference-cancelling types in use.

The design also features a two-wire system which has a central ground return path. It is possible that through the use of higher level lines some shielding previously essential in the original design could be eliminated. However, since interference tests have not as yet been conducted, sufficient shielding has been incorporated in the new design to insure adequate interference attenuation.

3.3.1.5 LIAISON SYSTEM

Airborne liaison systems are designed primarily to serve as an air-to-ground communication link utilizing either voice or code transmission.

In general, such systems consist of a transmitter, transmitter key, fixed antenna, radio receiver, antenna coupler, dynamotor, and interconnecting cables. The transmitter and receiver operate over a frequency range of from 200 to 500 kc and 1.5 to 18.0 mc. An audio output signal provided by the receiver is available at any one of the interphone stations. The installation of a liaison set in a typical aircraft has been selected as an example of an airborne liaison system. The discussion to follow applies specifically to this particular system. However, because of the similarity of all such systems, there will be no loss of generality.

A fixed antenna is mounted on the top or under the fuselage near the radioman's compartment. All tuning adjustments are made at the transmitter or receiver in the radioman's compartment. No remote controls are provided. A liaison receiver audio output and a liaison transmitter microphone input are provided at each interphone control box. In general, one interphone control box is located at each of the stations for the pilot, co-pilot, navigator, and radio operator. The power source

bus, antenna coupler, and Loran receiver are mounted in the compartment behind the pilot. There are no components located in the tail or wing sections. Interconnecting cords and cables are generally bundled with other electrical wiring passing through the compartment. Navigation and search radar controls, radio altimeter indicator, Loran indicator, etc., are examples of types of equipment whose wiring is frequently found in a typical bundle. Exciter regulators and radars are examples of equipments that would be mounted in the same aircraft sections.

Since in this system the receiver output is an audio signal in the headset, any unwanted signals which find their way into the receiver case must be capable either of producing audible interference in the headset or of preventing normal functioning of the receiver in some other way to constitute an interference problem. In complete systems such as this one the paths of entry of unwanted signals may be many and devious. In this case of the particular receiver in this representative system, the many paths of entry are illustrated in Figure 3.3.1.5-A. This figure shows also that while the functioning of the other components of the system may be unaffected by interference signals, they may serve as a part of a path of entry for the undesirable signals. For example, interference was picked up from the radar modulator on the lead between the dynamotor and the liaison transmitter which would not have occurred had this lead been routed differently. This interference did not affect the operation of the transmitter, but within the transmitter itself the dynamotor and antenna leads were bundled together and the interference reached the liaison receiver via the antenna connection. Here it was amplified and appeared as serious audio interference in the receiver output.

A brief explanation of the interference problems as shown in Figure 3.3.1.5-A follows, some solutions are shown in Figure 3.3.1.5-B but for the sake of clarity only a few are included.

(a) Interference was picked up directly on the open-wire antenna lead-in which was conducted to the liaison receiver causing unsatisfactory operation. It was determined that the antenna lead to the receiver could be shielded but that the transmitter antenna lead could not since the shielding capacitance would detune the transmitter output stage. The solution used is illustrated in Figure 3.3.1.5-B, and may be considered a satisfactory interim measure. Proper design, however, would permit the use of a shielded antenna lead-in for both the transmitter and receiver since harmonics as well as the carrier frequency of the transmitter are being radiated from the unshielded transmitter antenna lead-in and constitute a source of interference to other equipment.

(b) Interference from the radar modulator was picked up on the lead between the dynamotor and the liaison transmitter and was conducted into the transmitter. Within the transmitter the antenna lead to one of the transmit-receive relays was bundled in with the dynamotor and other power wiring. The interference was thereby coupled into the antenna lead to the receiver and was conducted to the liaison receiver causing unsatisfactory operation. Removing the receiver antenna lead and rerouting it as in (a) above and as shown in Figure 3.3.1.5-B is one possible solution. Another solution would be to isolate the antenna lead from the power wiring within the transmitter by shielding and/or rerouting. The latter would be preferable practice in initial design stages since it is poor practice in any event to bundle interference susceptible wiring with other wires likely to conduct radio interference. In other

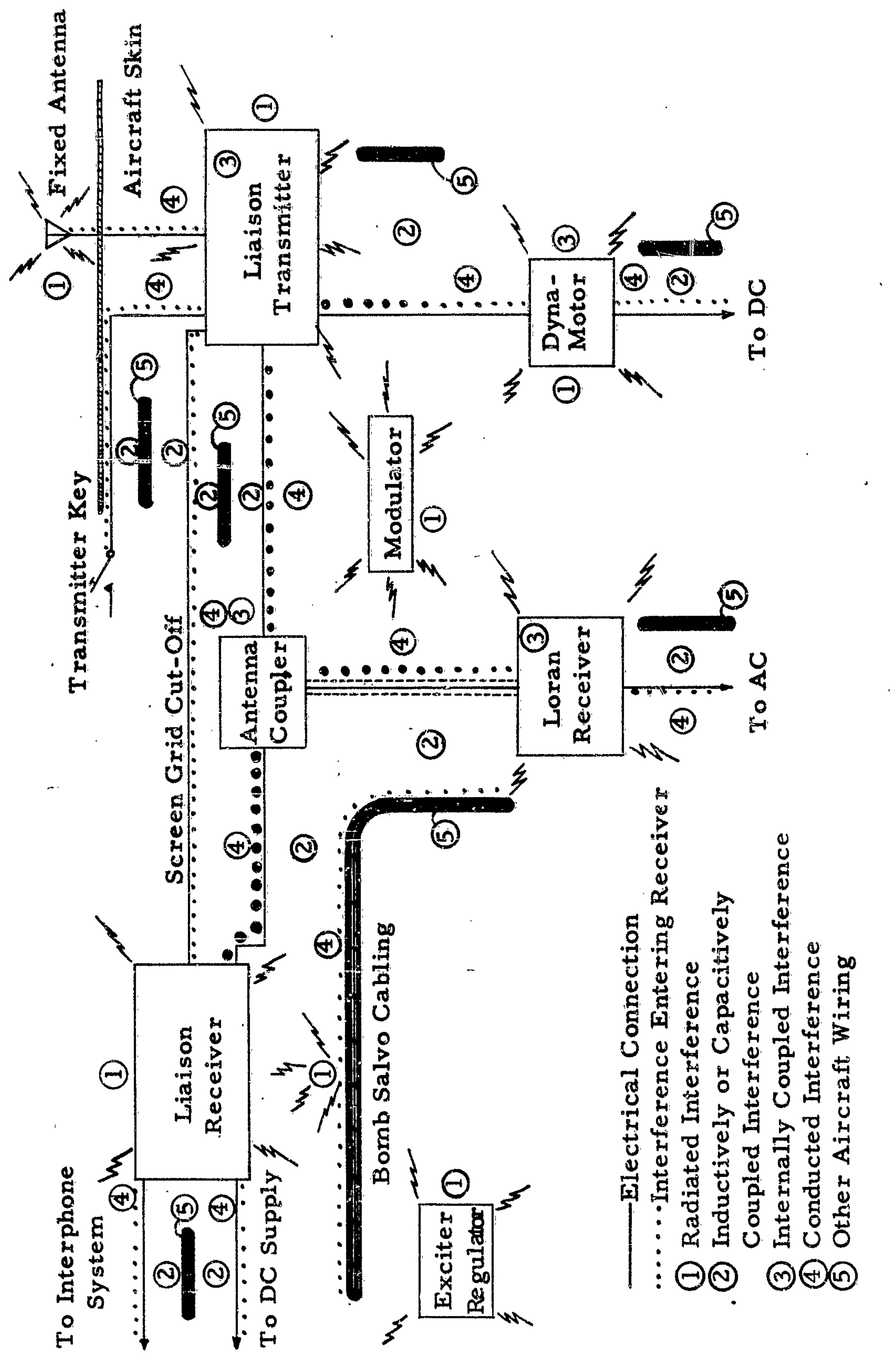


Fig. 3.3.1.5-A Paths of Interference Signals in a Liaison System

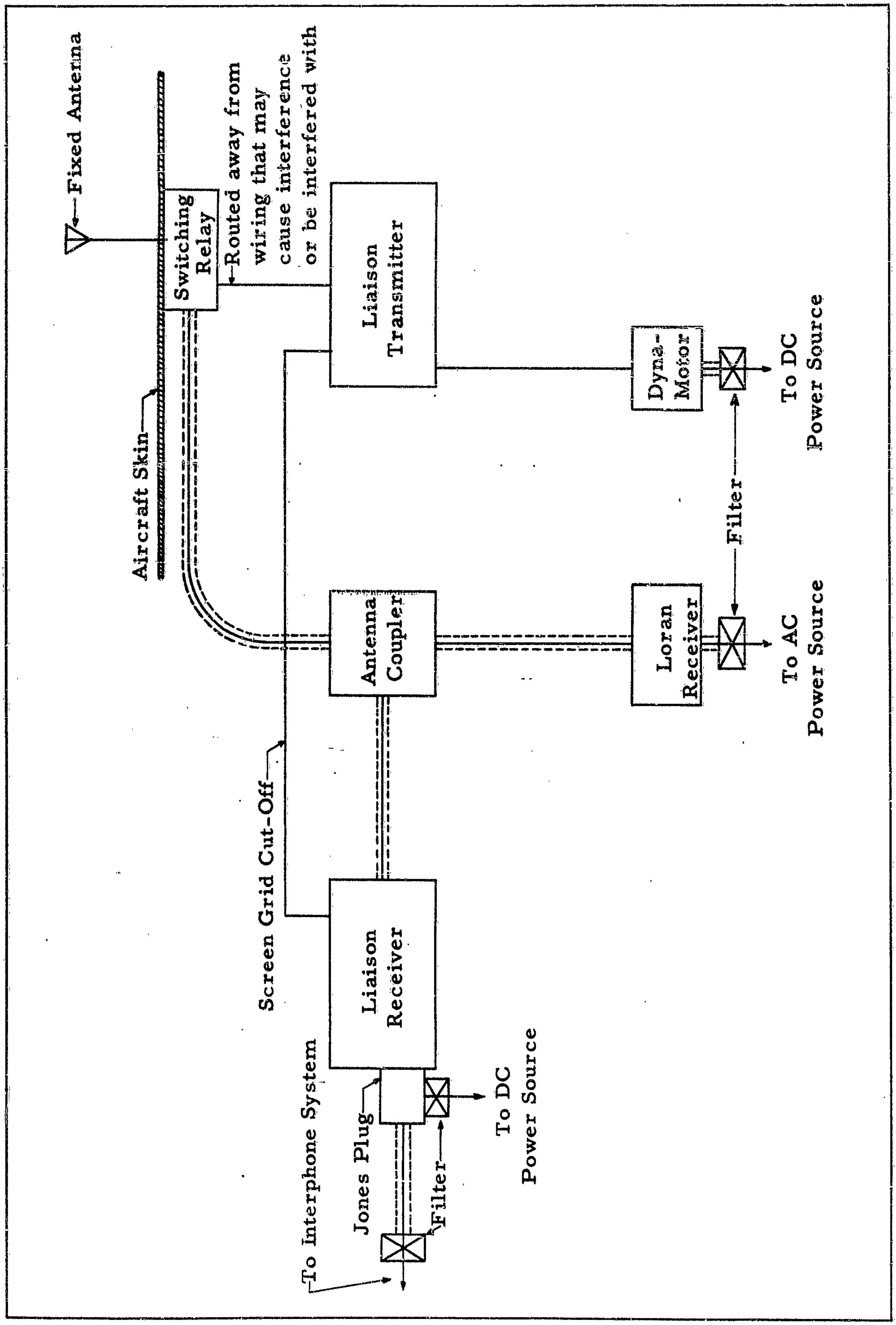


Fig. 3.3.1.5-B Modifications of the Liaison System to Minimize Interference

circumstances a filter in the lead from the dynamotor to the transmitter might be dictated, but only as a last resort since it would involve added equipment and weight.

(c) Interference was conducted into the receiver on the interphone and DC power leads, which caused unsatisfactory operation. At this stage of design, filters were installed in the troublesome leads as shown in Figure 3.3.1.5-B in order to reduce the interference to a tolerable level. This means added equipment, weight and cost. Preferably, the receiver should have been designed for less susceptibility to interference on these leads by improved shielding of the RF and/or IF stages and possibly better routing of these interphone and power leads within the receiver. Electrical cleanness (isolation) in the wiring of transmitters and receivers is preferable to mechanical neatness achieved by carefully bundling all the wires together into a cable and lacing them together.

(d) In this particular installation the bomb salvo cabling (an entirely independent system) was routed very close to the exciter regulator and to the liaison receiver cover in the region of the cooling louvres. Interference was coupled from the exciter to the bomb salvo cabling and thence through the louvres in the receiver cover to the internal wiring of the receiver resulting in objectionable audio output in the headsets. Rerouting of the bomb salvo cabling in the neighborhood of the exciter regulator to a point not less than six inches away resulted in interference-free operation. Several faulty design practices are involved in creating this particular problem:

1. The exciter regulator design was poor in that its generated interference was not suppressed properly.

2. Cables should never be run close to known interference sources whenever practicable.

3. The receiver case shielding properties were poor due to poor design. Wire mesh over the louvres and better bonding of the case to the frame are required.

(e) Interference was coupled from the radar modulator to the AC power lead of the Loran receiver and was conducted into the case of this unit. Inside this receiver the antenna lead was bundled in with the power wiring and interference was therefore introduced into the liaison receiver on the antenna lead. One solution is the addition of a filter to suppress the interference as shown in Figure 3.3.1.5-B. Proper design of the internal wiring of the Loran receiver to properly isolate or shield the antenna lead would be preferable since the filter would then be unnecessary.

3 3.1.6 LORAN SYSTEM

The Loran system is a navigational aid that enables the operator in an aircraft to fix his position over land or sea by means of the reception on his receiver-indicator of special radio signals from the ground installations.

The ground installations consist of groups of transmitting stations operating on the same radio carrier frequency which emit a steady succession of pulses in all directions. The stations operate in pairs, a master station triggering a slave station

by means of a radio link which synchronizes the pulses from the two stations.

The receiver-indicator receives and measures the timing of the pulsed signals and transcribes them to a visual indication on the Loran scope. With the aid of charts, tables, and tabulations the received signal can be interpreted and the location of the plane established.

The system operates in the high frequency range. The power supply is AC, 80 or 115 volts, 360 to 2460 cycles per second. In an aircraft the main units of the Loran system consist of an antenna, interconnecting cabling, either a passive coupler or a preamplifier coupler, and the receiver-indicator.

In a typical installation, the Loran system does not use its own antenna but couples onto the antenna of another installation. As a result unusually long runs of lead-in wire to the receiver-indicator are sometimes necessary. To compensate for the subsequent line-loss, a preamplifier is provided at the antenna coupler. When lead length is not excessive a passive coupler is used. The liaison system with its antenna approximates the needs of the Loran system and is the usual coupled system. In a heavy bomber installation, the Loran system uses the liaison antenna with an unshielded lead-in wire through a switching relay in the liaison transmitter to a preamplifier antenna coupler and a shielded lead-in wire from the antenna coupler to the receiver-indicator.

The Loran receiver, like any other receiver is subject to various sorts of interference. This interference may come from transmitters aboard the aircraft, aboard neighboring aircraft, from ground installations, or from deliberate enemy jamming. Typical interference patterns as seen on a scope would appear as shown in Figure 3. 3. 1. 6-B.

The Loran receiver has not proven to be a source of interference itself. The local oscillator is adaquately shielded, the case construction is apparently good. However, when operating as a system, the performance has not been entirely satisfactory.

The unshielded antenna lead-in prior to the coupling stage has proven to be an efficient coupling path for radio interference currents. See Figure 3.3.1.6-A. The preamplifier at the antenna coupler amplifies the interference which is then conducted through the shielded lead-in into the receiver. In most cases of this interference, the Liaison receiver is also affected. By the nature of the reception of the two receivers, a higher level of interference can be tolerated in the Loran than in the Liaison so subsequent filtering, shielding, etc., of the interfering components that cleared up the interference on the Liaison also cleared it up on the Loran. (See Paragraph 3. 3. 1. 5 on Liaison System.)

Poor interference-free design is apparent when the AC power lead-in plug is a few inches from the antenna lead-in plug. This provides a coupling path onto the antenna lead-in cable for interference conducted on the power line. These interference currents ultimately affect the liaison receiver (see Paragraph 3.3.1.5.on Liaison System) as well as appear as grass on the Loran scope. It was necessary to filter the AC line in order to eliminate this interference. Proper isolation in the original design would have eliminated the need for this filter.

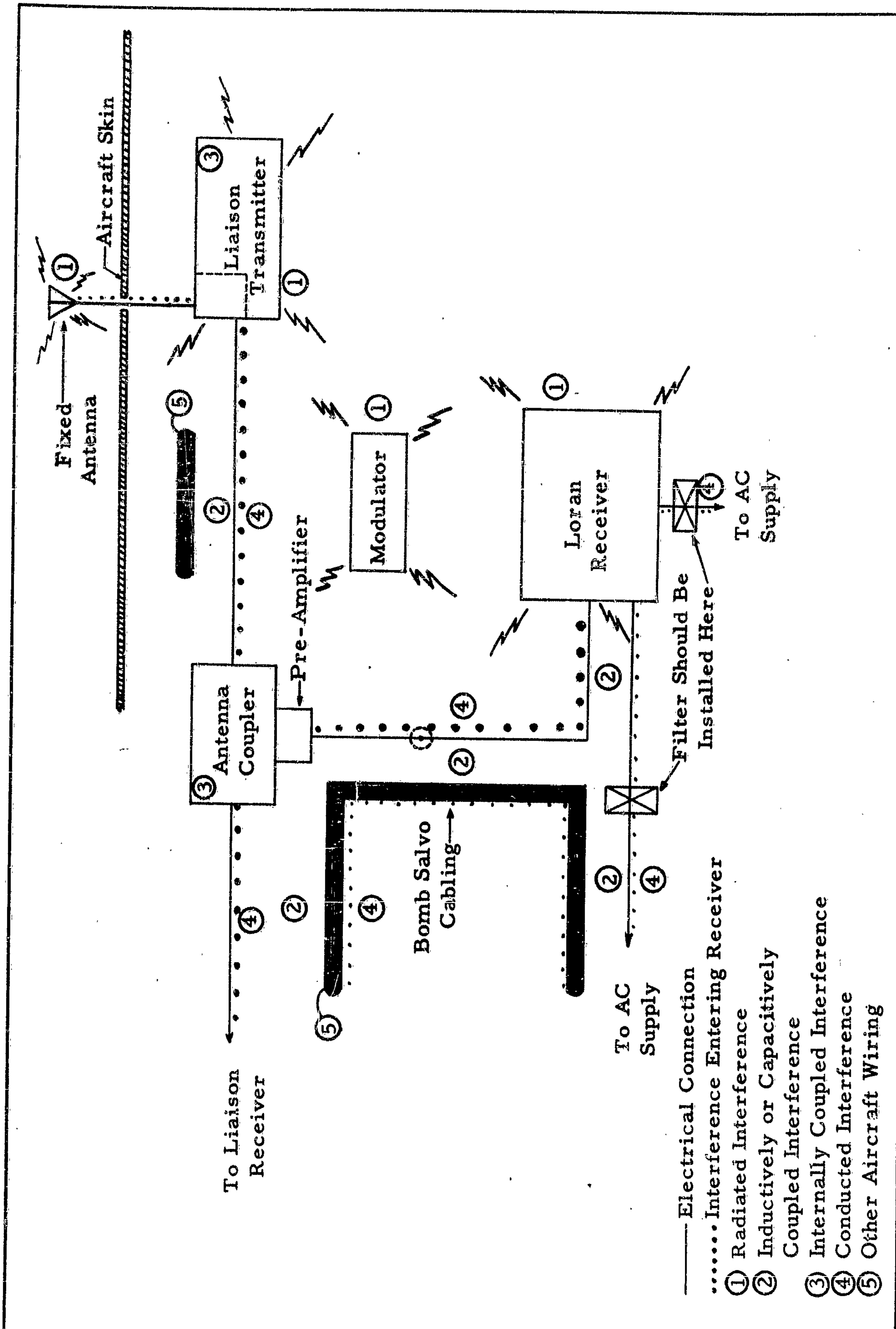


Fig. 3.3.1.6-A Paths of Interference Signals in a Typical Loran System

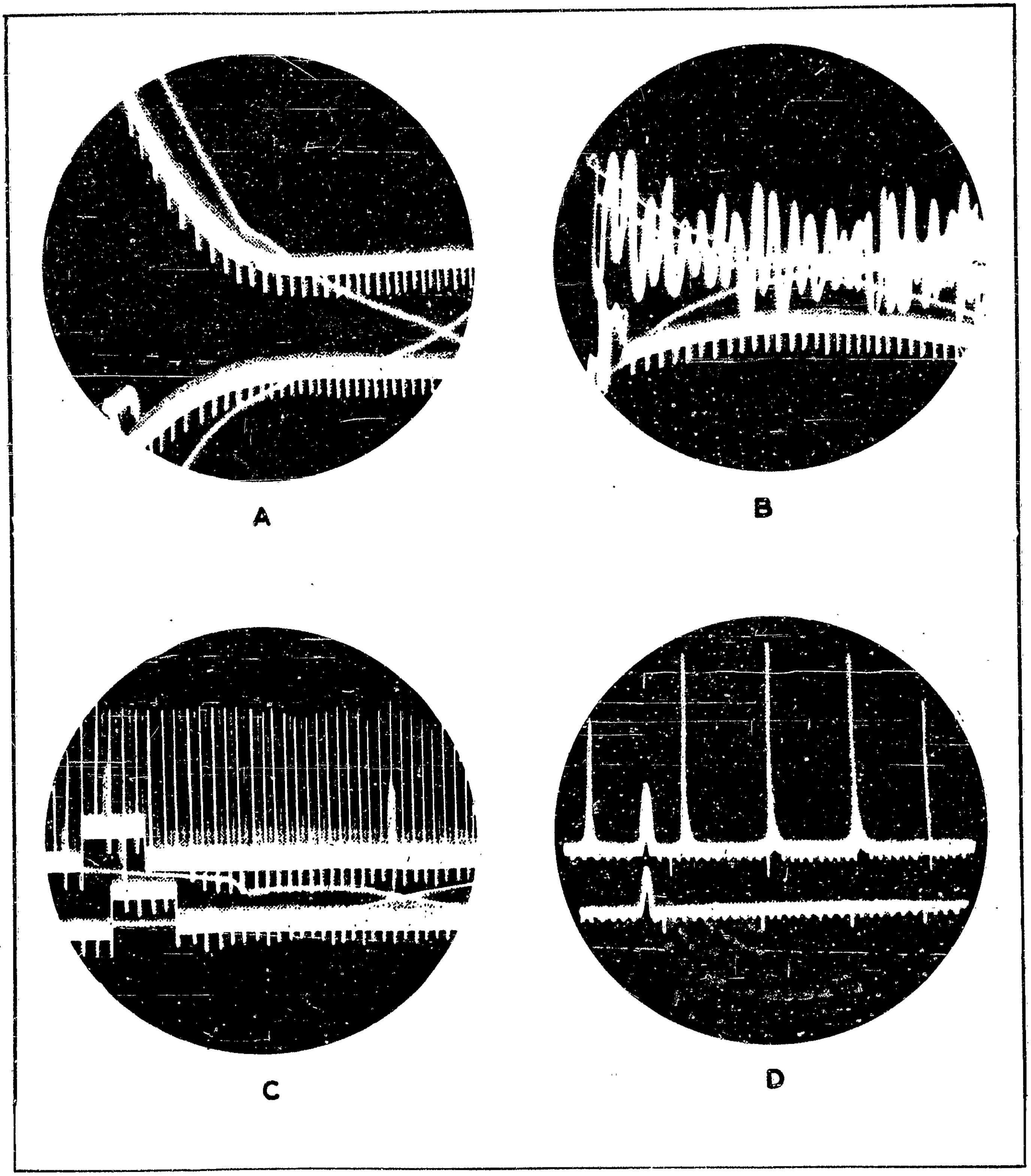


Fig. 3.3.1.6-B Typical Interference Patterns as seen on Loran Indicator
(A) CW or Radio Telegraph
(B) Radio Telephone
(C) Radar on Slow Sweep
(D) Radar on Fast Sweep

The preamplifier produces interference problems particular to this installation. The non-linear elements in the preamplifier make the system susceptible to cross modulation. A simple illustration of this would be: The Loran receiver is tuned to 2 mc; a 3 mc and a 1 mc signal can mix in the preamplifier stage and the result be admitted as a 2 mc signal in the receiver.

Another problem presented by the preamplifier stage is the system's susceptibility to overloading because of the wide pass-band. An unwanted signal can be amplified to a proportion which overloads the front end of the Loran. In a heavy bomber installation, rendezvous equipment was producing interference in the Loran which was extremely serious because of its high magnitude and the dangerous fact that the interfering signals could be mistaken for an actual Loran station. The interference was coming from a strong, pulsed, radar signal at about 200 mc. It was admitted through the Loran antenna, amplified at the coupler stage, overloaded the front end of the receiver, and gave a visual signal on the indicator scope. Preliminary investigations and tests indicated that a low-pass filter inserted in series with the Loran antenna, between the preamplifier and the antenna, should eliminate most of the interference. Further tests are necessary to determine the feasibility of such a filter, and if so, the most efficient arrangement considering insertion loss and matching, and terminating impedances. Systems using the passive coupler have not been subject to this type of interference.

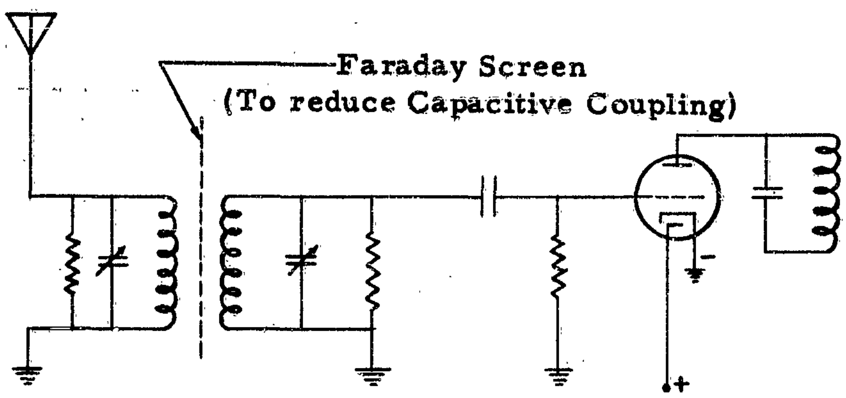


Fig. 3.3.1.6-C Tuning Device Installed
Ahead of Pre-Amplifier in Loran Receiver

Improved original design of the preamplifier would have effectively reduced the susceptibility of the system to both cross modulation and overloading. Consideration should have been given to the rejection of frequencies of undesired signals. A tuning device on the front end of the amplifier would be one means of increasing rejection. Figure 3.3.1.6-C shows such a tuning device. A band-pass filter with sharp cut-off characteristics covering the band of the Loran equipment as shown in Figure 3.3.1.6-D, would also give the necessary rejection characteristics.

The Loran system incorporates interference suppression techniques by providing for shielded lead-in cable, good case shielding against interference fields, and adequate shielding of the local oscillator to prevent interference with other equipment. Ideally, it should have its own antenna installation where it would be possible to minimize lead-in cable length; however, the increasing number of antennas in a

bomber installation makes it necessary to double up with another system. Good installation practice can minimize this length, but it may still be long enough in some installations to make a preamplifier stage necessary. Rejection of undesirable frequencies should be provided for in the preamplifier to minimize malfunctioning. Poor design is exemplified by the proximity of the AC power plug and antenna lead-in plug. Coupling of interfering currents has occurred between the power lead and the antenna lead and filtering has been necessary. Proper isolation would have eliminated the need for this filter. The front-end rejection of undesirable frequencies in the receiver could be improved.

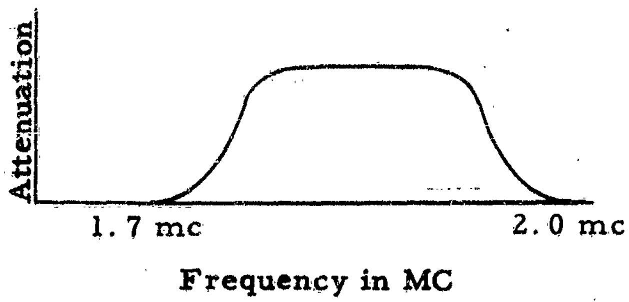


Fig. 3.3.1.6-D Attenuation Characteristic of Band-Pass Filter for Use With Loran Receiver

3.3.1.7 SHORAN SYSTEM

The Shoran Radio Set is a short range navigational system used in present day aircraft to determine the aircraft's position under instrument-flight conditions without visual reference to the surface of the earth or to celestial bodies. Two ground stations located a considerable distance apart are required in conjunction with the airborne installation in an operational system. During operation, pulses of amplitude-modulated radio waves of very high frequency are transmitted. Radiation from ground stations is moderately directional; radiation from the airplane is non-directional.

A typical Shoran system is shown in Figure 3.3.1.7. Two identical antennas and their bases (one for receiving and one for transmitting) are provided. The antennas must preferably be mounted on a large, relatively flat, metallic surface of the airplane. The particular location of each antenna depends on the type of airplane in which the equipment is installed and must be predetermined for each type by radiation field measurements. The antennas should preferably be shielded from each other by the fuselage or wings of the airplane. Also consideration should be given to the location in respect to the other receivers and transmitters in the airplane. In addition, attention is called to the fact that the transmission line (coaxial cable) which connects each antenna to the equipment enters the base of the antenna through a feed-through insulator and must be so laid out that the transmission lines make bends of no less than 4-inch radius to prevent damage to the lines.

The indicator is shock mounted in a frame which allows for a computer unit (when used). The receiver is installed within the indicator so that the controls of both units are accessible from the same position.

The indicator unit includes the circuits for generating the various sweep and

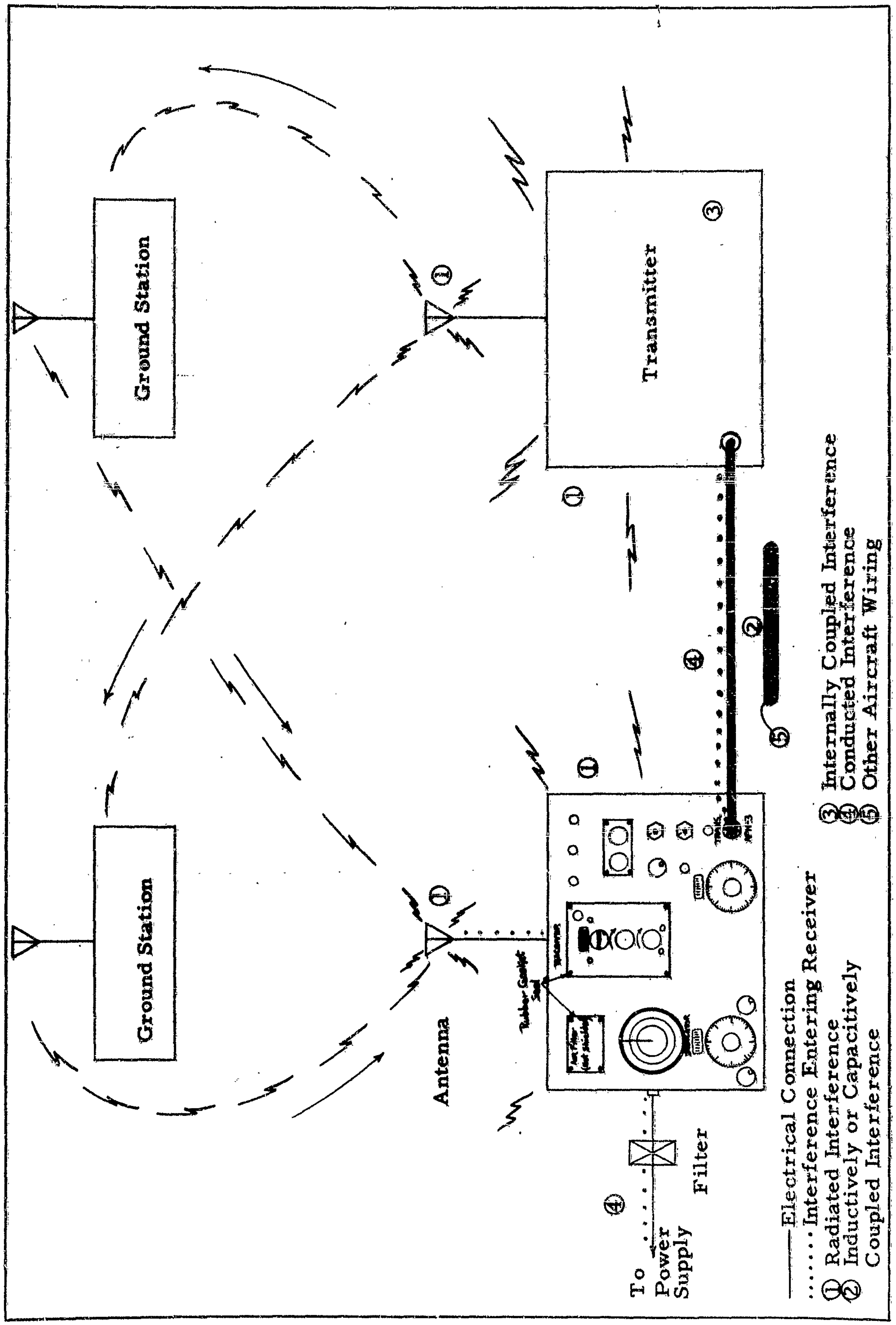


Fig. 3.3.1.7 Paths of Interference Signals in a Typical Shoran System

blanking voltages necessary for operation of the indicator tube. It also contains timing and phasing circuits, pulse-selector circuits, and keying-pulse-generating circuits for controlling the operation of the airplane transmitter. The indicator is cooled by air circulation which is provided by a blower motor within the unit with unscreened louvres over the air intake. The cathode ray indicator tube operates with a circular sweep and the signals, applied to an axial electrode within the tube, cause radial deflections of the electron beam. The signals thus appear as inner or outer pulses protruding from a circle on the screen of the indicator tube, and their relative positions serve to determine the position of the airplane.

The receiver utilizes a super-heterodyne circuit through which signals from the ground stations are amplified and converted into a video signal. This signal is then applied to the axial electrode of the indicator tube. The oscillator is the main source of interference. It radiates from the case, external wiring, and antenna.

The case is described as a dust cover and that is the only function it serves. There are rubber gaskets between the air filter and indicator case, the receiver and indicator case, and the connected sides of the indicator case. There is no screening on the front panel to prevent radiation both in and out at the various controls and other breaks in the case. The case is poorly constructed with loose tolerances and of non-rigid construction so that sides buckle in with little applied pressure. Tests in a shielded room using a probe showed high levels of oscillator radiation from the rubber-gasketed joints. Paint was scraped off the contact surfaces of these joints and shielding gasket braid was substituted for rubber gaskets. Shielding was applied to all apertures in the front panel so that contact with the control rods was made. Screening was placed over the air filter for shielding purposes. These measures effectively "bottled up" the oscillator radiation from the case.

All these paths of leakage described above also serve as paths of entry into the receiver case for radio interference. Difficulties have been encountered in installations where it was necessary to shut off other transmitting equipment in order to operate Shoran systems. Information as to coupling paths is not available and particular examples are classified material. It is assumed that the corrective measures taken to contain the oscillator radiation will also serve to keep out radio interference; however, tests were only conducted in a shielded room and it has not been proved in an actual installation.

Installation difficulties have been encountered with the Shoran transmitter in interfering with other receivers by both shock excitation and the generation of harmonics. The liaison receiver has been particularly susceptible, although other receivers have been affected by the Shoran transmitter. In a typical aircraft installation, the following modifications were necessary to produce a satisfactory system.

(a) The mounting of the liaison receiver was modified by the addition of grounding straps mounted inside of each shock absorber housing and by the removal of paint around the mounting bolts.

(b) The back end of the liaison receiver was filtered to prevent radio interference from Shoran transmitters and a wave trap was connected to the antenna binding post to prevent shock excitation.

(c) On the radio compass a choke coil was installed at the antenna binding post which acts as a low-pass filter to prevent shock excitation from Shoran transmitters.

(d) A capacitor was installed on the radio compass "look" antenna (reason not known).

Paths of entry into receivers have been both from the back and front ends. Filters have been installed in the back end and wave traps in the antenna leads. No information is available as to the coupling paths in any particular installation. Operating at a very high frequency and being pulse modulated, the Shoran transmitter has caused difficulties by overloading the receivers in the airplane and causing shock excitation. Most of the difficulty has been through antenna radiation although interference has been radiating from the case and from external wiring.

These difficulties can be overcome by incorporating the following features. Screening should be put across the air filter and conductive mesh gasket material installed instead of rubber gasket. All external wiring should be shielded to prevent radiation and possible coupling. The transmitter should be installed in the plane so as to take advantage of shielding by bulkheads, etc. Wiring should be routed to prevent coupling. Rejection of receivers can be improved by adding wave traps to antenna lead-ins and by filtering the back end when feasible.

In a shielded room with Shoran transmitter and liaison receiver antennas set 6 feet apart, a high level of interference was recorded in the VHF and UHF frequency range. The liaison receiver was picking up harmonics of the fundamental with spurious radiations, harmonics of the pulse repetition rate with spurious radiations, and shock excitation from the fundamental. Better internal circuit design is necessary to prevent the generation of harmonics (see Paragraph 3.3.1.2 on VHF System). Specific examples of bad practice in circuit design of Shoran transmitters are not available.

3.3.1.8 NAVIGATIONAL RADAR SET

A navigational radar set is a dual-purpose equipment designed to operate by switch selection either as a radar beacon or an interrogator-responder. When used as an interrogator-responder, its function is to guide the airplane within 200 yards of the beacon location. As a beacon, it responds to other interrogator-responder radars within its frequency range. As shown in a typical wiring diagram, Figure 3.3.1.8, a representative set would consist of: (a) receiver-transmitter, (b) indicator, (c) control box (power), (d) control box (frequency), (e) video gate, (f) transmitting antenna, (g) receiving antennas, (h) reflector antennas, and (i) mountings, cables, interconnecting plugs, etc.

The functional operation of a typical radar set when used as an interrogator-responder includes the following. The pulse-generating circuits furnish the modulator with pulses at various rates which may or may not be synchronized with associated radar equipment. The pulse generating circuits also furnish a synchronizing pulse for the indicator and a suppression pulse for the IFF equipment. The pulses from the modulator cause the transmitting oscillator to oscillate for four to six microseconds at the pulse repetition rate. This radio-frequency energy is fed to the transmitting antenna and the transmitted energy is received by an associated transponder,

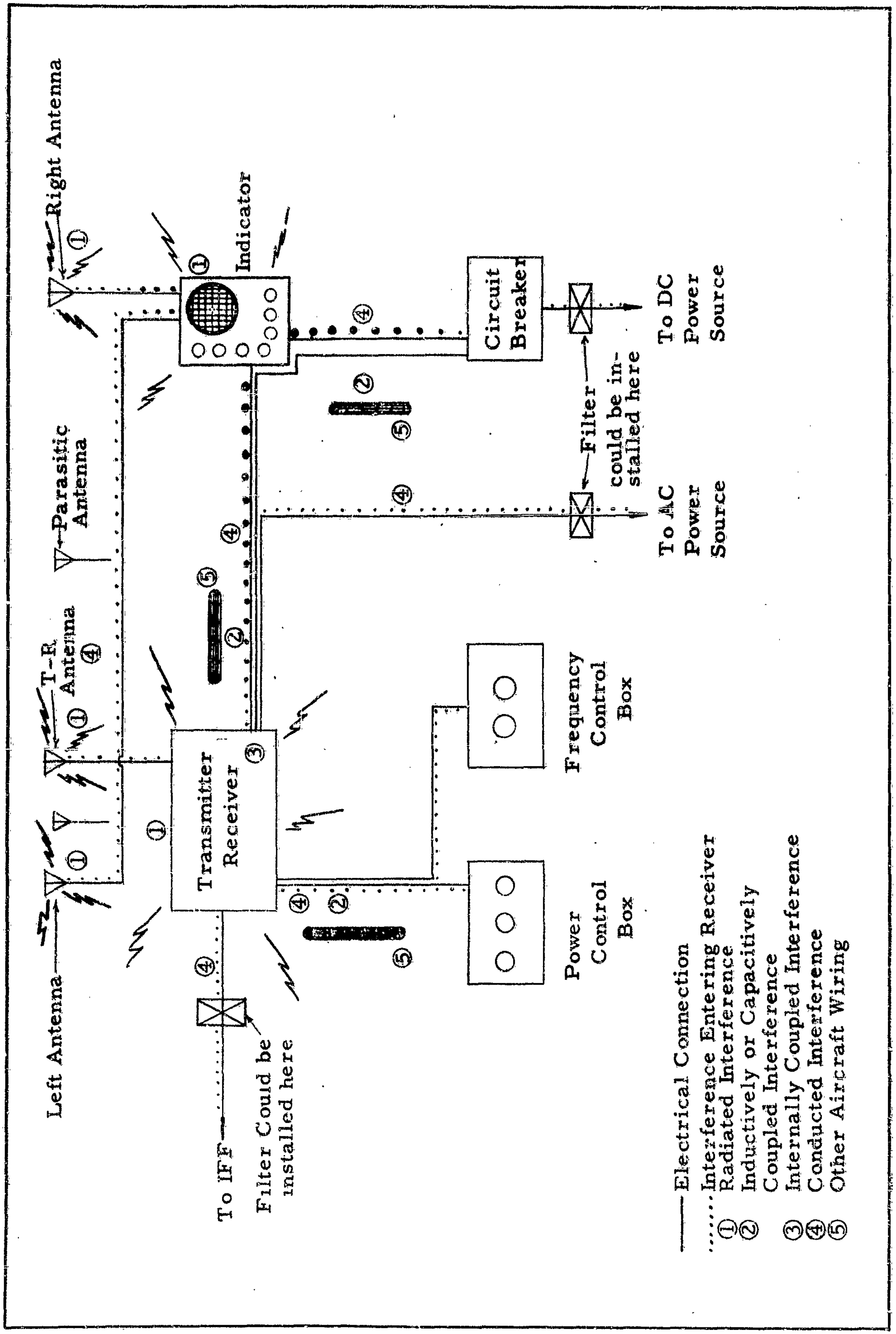


Fig. 3.3.1.8 Paths of Interference Signals in a Typical Radar Navigation System

which transmits a radio-frequency reply. The reply is picked up by the receiving antennas, the outputs of which are alternately fed into the receiver section by the antenna switch located in the indicator. The video output of the receiver is synchronously switched with the receiving antenna in the indicator and applied to the horizontal plates of the cathode-ray tube in the indicator.

When the radar set is used as a transponder (beacon), the video output is connected to the pulse-generating circuit. Thus the transmitter is triggered each time a pulse is received from the associated interrogator-responder.

The set operates in the VHF range on any one of eight preset frequencies. The receiver-transmitter unit is designed to maintain a high degree of frequency stability; the frequency drift of both components is less than one megacycle between ambient temperatures, -55° C. and +71° C. Power is obtained from a 115-volt, 400-1600 cps supply. In addition, a 24-28 volts, 2-ampere, DC source is required.

The transmitting antenna is usually installed on the underside of the nose of the airplane and, when possible, in the center of a 36-inch diameter, flat surface. Best results can be achieved if the antenna is within ten degrees of vertical to the airplane's line of level flight.

Receiving antennas are installed on each side of the airplane, symmetrical about the line of flight. The exact location for best results would have to be determined experimentally because it would differ with each model of plane but generally the two antennas are installed on the underside of the wing, outboard from a vertical reflecting surface. Installation of the indicator and receiver-transmitter in the plane should be convenient to the operator while sitting in a comfortable position to view the display tube. A minimum of wiring length and isolation from sensitive receivers are also installation considerations.

Coaxial cable is used from the receiver-transmitter to the antennas and from the receiver-transmitter to the indicator. Power leads and control wiring are unshielded. As installed in a typical airplane, an analysis revealed that the internal circuitry of the receiver-transmitter was not originally designed to suppress modulator interference. Wires carrying interference were coupled with "clean" leads and the cooling blower motor was generating undesirable voltages which interfered with the receiver-transmitter as well as with other equipment. Harmonics and overloading from pulses coming from antenna radiation affected other receivers. Interference fields were leaking out of the case and getting to the power and control wiring which then provided a coupling path to any adjacent wiring. The video and antenna leads were shielded according to good design practice; however, neither the AC nor DC power leads were filtered and hence constituted a source of interference. Also the lobe-switch drive motor was found to be an interference generator. While not a source of interference in itself, the interconnecting wiring provided a coupling path for interference generated by the receiver-transmitter.

In order to suppress interference from this typical navigational radar set numerous modifications were necessary. The following list of modifications were recommended by the manufacturer:

(a) The blower motor was changed from DC to AC to eliminate triggering of

the beacon function by brush-spark interference and to prevent interference in other equipment.

(b) The receptacles for power and control wiring were changed to a new type having ceramic by-pass condensers integral with the contact-pin construction to reduce leakage from the receiver oscillator and other interference sources in the unit.

(c) A shield was designed and installed on the radio-frequency tuner chassis of the receiver to enclose and confine the field of the line-type oscillator. Since it was not considered in the original design, the shield had to be modified to provide clearance holes for turret-wheel contacts and the openings used for access to trimmer condensers had to be capped. This added shielding materially reduced oscillator leakage and also suppressed the oscillations radiating from the radio-frequency stage which had occurred in the original design.

(d) Additional bonding was necessary for the RF tuner chassis to reduce chassis potential and leakage. It was also necessary to install chokes, by-pass condensers, and filter resistors both in and outside of the RF tuner to reduce leakage and radiated interference. A series resistor was installed in the mixer-plate output lead to reduce the oscillator leakage through the IF stages into the circuit wiring and radiation from the receiver.

(e) A small amount of delay bias on the video detector diode reduced the interference due to triggering of the beacon.

(f) The screen resistor of the RF amplifier tube was lowered from 33 to 10 kilohms to increase stage gain and improve the signal to interference ratio of the receiver.

(g) The tuning-slug screw in the oscillator-vernier mounting block was found to move with normal vibration and caused both amplitude and frequency modulation of the oscillator. A detent device was incorporated to prevent this vibration and correct the difficulty.

(h) Decoupling resistors were provided in the suppressor input and output circuits so that interference leakage along interconnecting cables would not cause faulty operation of paired units in beacon and interrogator-responder service.

(i) The receiver-transmitter plate-supply voltage was lowered to approximately 265 volts, which is 30 volts less than that used in the original design. The loss of receiver gain which accompanied this change was made up by lowering the cathode resistor of the fifth IF stage from 470 to 270 ohms. The chief advantage of this change was to lower the plate dissipation in all the tubes, especially in the 6AK5 of the RF stage.

Lowered plate voltage for the transmitter also was intended to eliminate voltage breakdown and reduce corona on high voltage leads and terminals at high altitudes, a condition which had made the triggering due to interference so severe that

the beacon function was not usable. Transmitter-power output indication was satisfactory at the lowered plate voltage provided a tube with ample filament emission was used in the power-output indicator device. In addition, several detailed changes were made to increase air-gap length and decrease corona on the high-voltage, power-supply layout. A recent investigation on a heavy bomber installation revealed that the navigational radar was producing interference in the command receiver, the radio compass, the liaison receiver, the VHF receiver, and the Loran receiver. The interference produced in the Loran was particularly serious because of its high magnitude and the dangerous fact that it can easily be mistaken for an actual Loran station.

Preliminary tests indicated that the interference was being radiated from the radar transmitting antenna and being picked up by the antennas of the above receivers. Overloading was occurring and the interfering signal was being admitted. A low-pass filter was inserted in series with the Loran antenna and it appeared that most of the interfering currents were eliminated. Further investigations have to be made to determine the feasibility of this type of filter as well as to consider the insertion loss and the matching and terminating impedances.

The above discussion should serve as an excellent example of what is likely to happen if proper design practices are not made an integral part of all phases of design. While it may be admitted that this system was made operable by making certain modifications, it must be understood that fundamentally "fixes" are inefficient and costly in terms of time, weight, and space.

3.3.1.9 RADIO ALTIMETER SYSTEM

Airborne radio altimeter systems are designed to indicate the altitude of the aircraft above the terrain by accurately measuring the time required for a radio signal to travel to the earth's surface and return, and interpreting this time interval in terms of distance in feet above the reflecting surface directly under the aircraft.

Representative equipment is designed to emit a frequency-modulated radio wave in a downward direction from the transmitter antenna. For both the low and high ranges, 0 - 400 feet and 400 - 4000 feet respectively, the carrier signal is in the UHF range. The earth's surface reflects a portion of this radiated energy, and this is received on a separate receiver antenna. During the time interval required for the signal to travel to earth and return, the frequency of the transmitter will have changed. A small fraction of the transmitter signal is transmitted directly to the receiver and mixed with the reflected signal from the earth. Since the transmitter is continuously changing its frequency, there will be a difference in frequency between the direct and reflected signal at any instant of time and this difference will be proportional to the distance of the aircraft above ground. The detector circuit converts this difference in frequency to a direct current which is indicated by a meter calibrated directly in feet of altitude above the ground. The detected current also operates a pair of relays which are preset to energize altitude-limit indicators.

In general, radio altimeter systems are composed of (1) transmitter antenna and receiver antenna, (2) a transmitter-receiver, (3) indicators, (4) limit-switch assembly, (5) limit-light indicator, and (6) inter-connecting cables. The installation of a representative radio altimeter set in a typical aircraft has been selected as an

example for discussion of design techniques for interference-free operation.

The transmitter and receiver antennas used on the radio atlimeter are identical and interchangeable. Each antenna consists of a one-half wavelength dipole on a suitable mounting structure with provisions for connecting a 50-ohm transmission line through a coaxial elbow adapter. The antennas are mounted on a suitable metallic surface which acts as a reflector. The antenna system is designed to radiate and receive maximum signal strength in a generally downward direction with a minimum transfer of energy from the transmitting antenna to the receiving antenna, and to minimize variations in the received reflected signal while executing a reasonable dive, climb, or banking maneuvers. The pilot's indicator, limit and marker lights are mounted within reach of the pilot when seated at the flight controls. There is no audio output for the radio altimeter - all indications are visual.

The transmitter-receiver unit is usually located as close to the antenna installations as practicable and consequently appears in the compartment nearest the antennas. Interconnecting cords and cables are generally bundled with other electrical wiring passing through the compartment. Navigational radar receiver-transmitter controls, search radar synchronizer, navigational radar indicator, Loran indicator, and radio altimeter indicator, are examples of types of wiring that are frequently found in a typical bundle.

Radio interference signals have gained access to the receiver circuits over paths of entry, as shown in Figure 3.3.1.9, and caused unsatisfactory operation of the radio altimeter system. Interference signals generated by the system may be introduced into other aircraft wiring systems over the same routes. In the radio altimeter system only the transmitter-receiver combination need be considered as a potential source of interference. It is also the only component of the system that is susceptible to interference.

In general, interfering signals can enter the transmitter-receiver case through the following paths: (a) transmitting-antenna lead, (b) receiving-antenna lead, (c) automatic-pilot altitude-control cable, (d) limit-switch cable, (e) indicator cable, and (f) the power-input cable. Interfering signals can be introduced into the receiver over any of these paths by conduction along the cables or through inductive or capacitive coupling between other aircraft wiring and the system wiring.

Interference problems in the radio altimeter system of the typical aircraft installation selected for discussion were caused by the following: (a) loose connections in the antenna transmission system due to the fact that plugs, adapters, receptacles, etc., were not tight or properly meshed and (b) field modulation which resulted from "secondary feed-through" paths having variable elements and were capable of effectively modulating the received signal.

As an example of an unusual type of interference which may affect radio altimeter systems, consider the reflecting surface of a propeller which is moving through the common field of the two antennas. The intermittent reflected signal thus picked up by the receiving antenna combines with and effectively modulates the desired signal reflected by the earth. If this field modulation, after passing through the detector and audio-frequency amplifier stages is of sufficient amplitude to operate the square-wave limiter, counts will be added to the limiter output signal and an excessive altitude

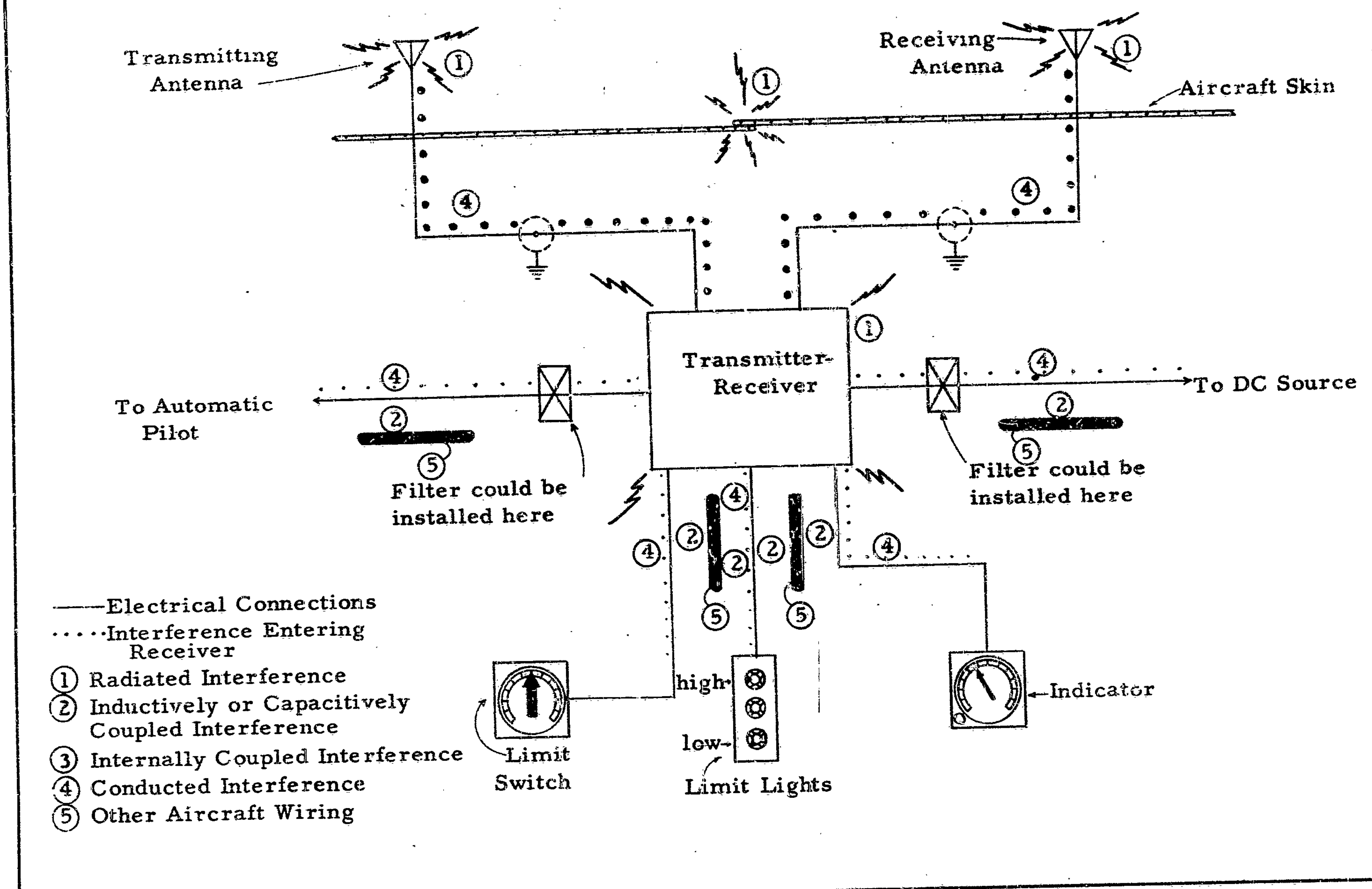


Fig. 3.3.1.9 Paths of Interference Signals in a Typical Radio Altimeter System

will be indicated.

A similar but more erratic effect is produced if there is a loose metallic part of the aircraft within the fields of the antennas which can shift or vibrate so as to cause an intermittent or rubbing contact with another metal part or surface. In this case, induced currents, set up in one of the surfaces by the transmitted signal, are modified both in phase and in amplitude by intermittent contact with the other surface similarly charged. Thus, transient currents, covering a wide band of frequencies, are produced in the field of the receiver antenna. Usual sources of such trouble are bomb bay doors, loose access plates, loose wheel fairing, trailing wire antennas, etc. When a source of field modulation is found, an attempt should be made to correct the condition either by electrically bonding or by completely insulating the two surfaces which are causing the trouble. If such correction is not feasible, it may be necessary to relocate the antennas so as to remove the source from the common field of the antennas.

Difficulties have been encountered in a typical aircraft installation where audio ripple originating in a motor has been conducted by the power leads into the transmitter and receiver affecting the altitude indicator by as much as 300 feet on the high range scale. The altitude indicator is in a particularly susceptible position because of the circuit arrangement. In a controlled situation, an audio oscillator could vary readings by 300 feet with little difficulty. In a typical installation, a 250 μfd capacitor was installed at an aileron booster motor in order to suppre , the audio ripple that was getting in through power leads and affecting the indicator. Proper transmitter-receiver design should have made the unit insusceptible to interference on these leads.

Interference in radio altimeters often manifests itself by (1) erratic altitude indications observed as jumps of 200 - 400 feet when flying over land at altitudes from 400 - 3000 feet, and (2) marked increase in altitude indication when switching over the high range while in flight at altitudes near 400 feet. It has been found that these effects are usually caused by electrical interference which effectively adds counts to the audio-frequency signal output of the detector. Such conditions are most likely to be encountered when the reflected signal is relatively low in amplitude as when flying over dry land.

The difficulties encountered with this system as explained above and as shown in Figure 3.3.1.9 could have been avoided had proper design and installation techniques been adhered to from the start. Two types of problems can be shown to exist by inspection of the figure. One group concerns the equipment designer and may be outlined as follows:

(a) The antenna lead-ins may be considered potential sources of interference to the system itself. Proper design will permit both to be completely shielded from the antennas to the transmitter - receiver, including correctly matched input and output circuits.

(b) Cabling between the transmitter-receiver and other units to which it must be connected may be considered similarly. Proper design of the internal circuits of the transmitter-receiver will insure that they are not susceptible to interference picked up on the cables. Additionally, the design should

be such that interference will not be conducted out of the transmitter-receiver on these cables. If a signal essential to the system operation is to be conducted betwe. units of the system and which might cause interference in other equipm ent, the signal should be contained within the cable by shielding.

The second group of problems concerns the installation designer and may be outlined as follows:

(a) The antennas should be installed outside interference fields of other equipment and antennas.

(b) Install the antennas so that the receiving antenna is shielded from the transmitting antenna.

(c) Install the other components of the system in as close proximity as possible using proper interconnecting cabling, and where practicable, away from strong interference sources and fields.

3.3.1.10 RADIO COMPASS SYSTEM

Airborne radio compass systems in use today are basically similar in function and design. The system consists of a radio receiver using a superheterodyne circuit and certain additional circuits necessary for radio compass operation. Two remote control boxes and two indicators permit operation of the compass from either of two separate positions in the aircraft. A relay unit including an autotransformer is provided to permit switching from one control box to the other. A vertical rod, non-directional antenna and a center-tapped, loop antenna together with the necessary interconnecting cords and power cable complete the system.

In general, radio compass systems operate over a frequency range of approximately 100 to 2000 kilocycles and are capable of providing:

(a) an automatic visual bearing indication of the direction of arrival of radio frequency signals,

(b) aural reception of modulated radio-frequency energy using either a non-directional or loop antenna, and

(c) aural-null directional indications of the arrival of modulated radio-frequency signals using a loop antenna.

The installation of a representative radio compass set in a typical aircraft has been selected as an example. The discussion to follow applies specifically to this particular system. However, because of the similarity of all such systems, no sacrifice of generality is suffered.

Loop and sense antennas are mounted outside the aircraft on the top or under the fuselage near the pilot and navigator compartments. The pilot's control panel and indicator are mounted within reach of the pilot when seated at the flight controls. There is another remote control panel and indicator mounted in the navigator's

compartment. A radio-compass audio output is provided at each interphone control box. In general, there would be a pilot, co-pilot, navigator, and radio operator interphone control box, each located within easy reach of the crew member. The relay, power shield, and receiver are mounted in the compartment behind the pilot. An inverter in the same compartment provides a 400 cycle per second, AC supply to the radio compass system. There are no components located in the tail section or wing sections. Interconnecting cords and cables are generally bundled with other electrical wiring passing through the compartment. Radar sets, computers, and auto-pilot trays are frequently mounted in the same compartment with the radio compass receiver.

Radio interference signals have caused unsatisfactory operation of the radio compass system. Visual bearing indications are particularly susceptible to interfering signals due to the fact that unwanted voltages appearing on the grids of the thyratrons, which position the loop antenna, can override the desired signal and cause erroneous or unstable visual indications. Aural reception and aural-null indications may become unintelligible when interfering signals of sufficient nuisance value are allowed to produce audible sounds in the headset.

There are two components in the radio compass system that must be considered as potential interference generators: (1) the drive motor in the loop antenna, and (2) the receiver itself. Design considerations for these components are discussed in Paragraph 3.2.1.1 and 3.4 respectively. The receiver requires particular attention because of its local oscillator and thyratron output.

An analysis of each component in the compass system reveals that only the receiver is susceptible to radio interference. Even though the functioning of the other components of the system is unaffected by interference signals, they may serve as paths of entry for undesired signals.

In general, interfering signals can enter the receiver case itself, through any one of the following paths, as shown in Figure 3.3.1.10:

(a) Separate inverter AC power supply cable,
(b) DC power supply cable,
(c) Loop and non-directional antenna cables,
(d) Flexible tuning shafts,
(e) Marker beacon receiving equipment, 14 to 28 and 220 volt DC supply cable,
(f) Remote control connecting cables,
(g) Headphone output cord, and
(h) Penetration of the receiver case or shield.

Interfering signals may enter the radio compass receiver by conduction along the cables as a result of inductive or capacitive coupling links between any of the system cables and other aircraft wiring. Such cable connections are those between the control boxes and the relay, relay and marker beacon receiver, relay and AC-DC power shield, relay and loop antenna, and indirectly, those from the loop-antenna motor control circuits.

Unwanted signals can also be introduced into the receiver through any one or more of the five openings in the metal receiver case. These openings are provided

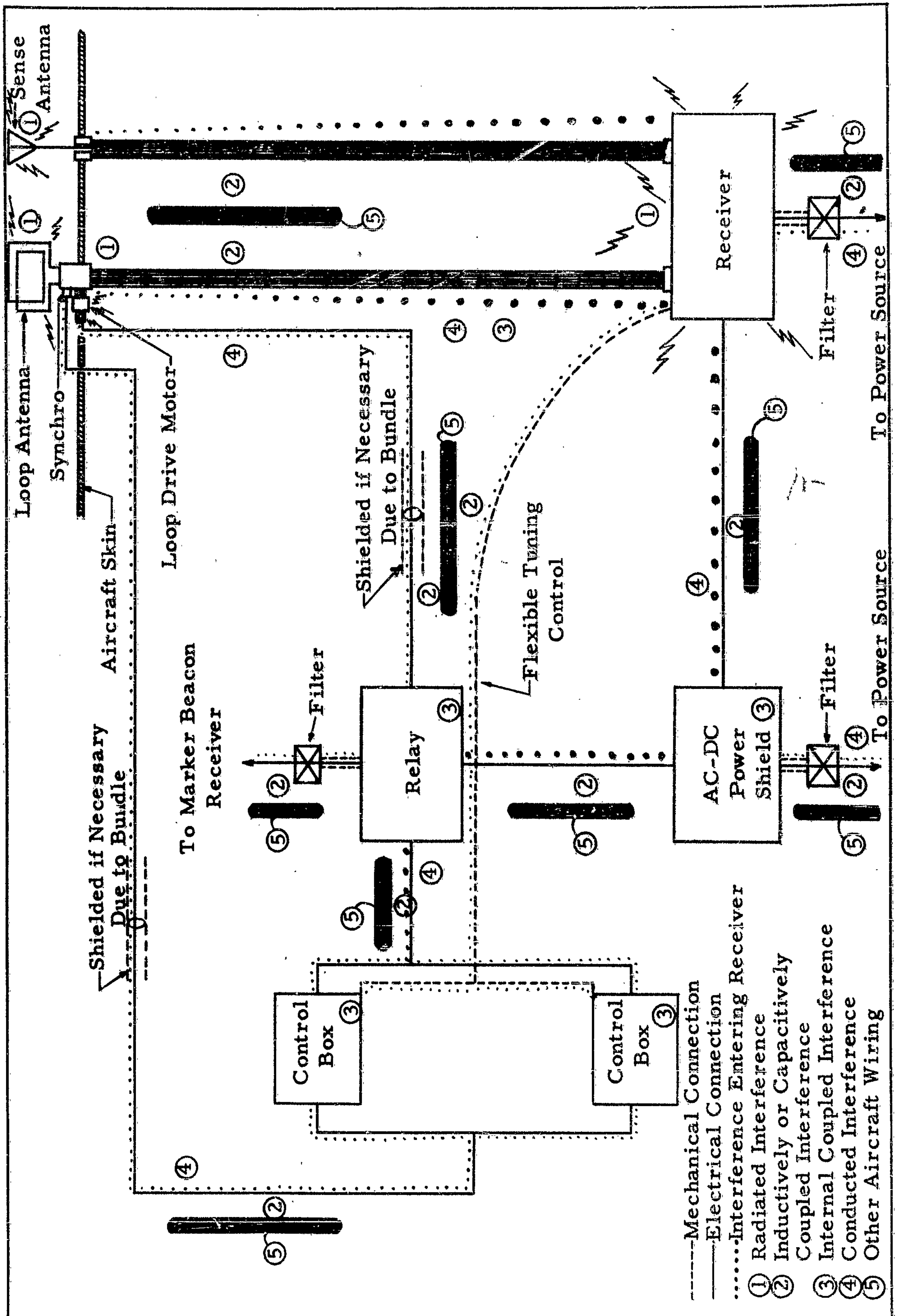


Fig. 3.3.1.10 Paths of Interference Signals in a Typical Radio Compass System

for the loop antenna and sense antenna leads, the AC-DC power cable, the headset or interphone audio-output cord, and the flexible tuning linkages. Interfering signals can be inductively or capacitively coupled into these cables directly and gain access to the receiver case. In a typical radio compass installation, an auto-pilot tray was located close to the receiver and the power cables of the auto-pilot tray paralleled the antenna leads and power cable of the receiver. This difficulty was increased by the use of an unshielded sense-antenna wire to the radio compass receiver. Interference was reduced by substituting a shielded coaxial cable for the unshielded wire, rerouting the loop cable away from the auto-pilot cabling, and braid shielding the entire cable bundle. These steps could have been taken in the original installation; however, the offending source, namely, the auto-pilot, should have been rendered interference-free by the manufacturer in the original design, by application of techniques outlined in Section I. Observance of good design techniques would have obviated the necessity for the preceding fixes.

Figure 3.3.1.10 illustrates the most vulnerable points for interference signals to enter the radio compass system. Unwanted signals picked up by the marker beacon antenna can reach the receiver through the marker beacon receiver and relay to the AC - DC power shield. From the power shield, a direct path is provided over the power cables to the receiver. Disturbing signals picked up by the loop or sense antennas have a direct path into the receiver case over the antenna wiring.

The degree of freedom from interference attained in any radio compass system will depend upon the care of the designer in applying radio interference elimination techniques, as discussed in other paragraphs of this book. The major interference in radio compass systems is due to conducted signals over the system wiring and also to inductively or capacitively coupled signals into the system wiring. In radio compass systems, particular attention must be paid to the filtering, shielding, and bonding necessary to avoid conducted and radiated interference from neighboring systems operating from the same power supplies and in close proximity. An effort should be made to avoid bundling the system cables with other cables that are likely to couple unwanted signals into the compass system. Also, the components of the system, as well as the wiring, should be located and routed so as to avoid proximity to obvious or probable interference generators.

Components should be placed and circuitry arranged in such order as to result in a minimum susceptibility to interference. Reference to Figure 3. 3. 1. 10 shows that the compass system wiring is directly connected to other electrical systems in the aircraft through three connecting cables, i. e., (1) receiver audio output to interphone system, (2) relay power supply to marker beacon receiver, (3) AC-DC power supply to power shield. These lines can be filtered to prevent interference from entering or leaving the system by conduction. Details of filtering and filter design are discussed in Paragraph 3. 1. 1 and Appendix VII. Proper design techniques, as outlined in other sections of the book, however, should have been applied in the initial design so as to make the equipment insusceptible to interference on these leads. There is also the possibility that the components of the radio compass system may not be at the same ground potential for a certain band of interference frequencies. Such a condition frequently occurs, due to varying impedance-to-ground characteristics for the circuit components, and consequently provides a path for circulatory interfering currents. Proper bonding will prevent these conditions, as pointed out in Paragraph 3.1.3. Each component should be adaquately bonded, and each shielded

cable should be bonded, not only at its terminal points, but also at each supporting strap, in the case of long cables.

Loop antenna and sense antenna leads should be electrostatically shielded. The synchro connections from the loop antenna to the pilot and navigator indicator are generally bundled together with other aircraft wiring and provide a means of coupling interference from the loop drive motor into other aircraft electrical and electronic systems. Consideration must be given, also, to the possibility of coupling interference from the other systems into the receiver through the loop antenna lead-in. If these cables are bundled with other susceptible-system wiring, or other probable interference-generating system wiring, they should be shielded, unless rerouting is practicable. The example, discussed above, where the auto-pilot caused considerable interference in the compass system, is a good illustration of the difficulties that can result when proper shielding techniques are not observed. Basically, however, the auto-pilot is the offender and source suppression of interference was not properly considered in its design.

3.3.1.11 RADAR FIRE-CONTROL SYSTEM

Radar fire-control and gun-laying radar systems (see Paragraph 3.3.1.12) have essentially the same components and serve basically the same function. These two representative systems were chosen to contrast design techniques pertinent to radio interference.

The radar fire-control system chosen is an all-weather system designed for high altitude interceptors that employ fixed armament. It has two primary functions, search and attack. In two place aircraft a transitional function, hand-tracking, is provided for the radar operator. In the search function, the fire-control system detects the enemy aircraft. Once the target has been located, the operator directs the antenna toward the target by hand-tracking and places the equipment on automatic track. At this time, the pilot's indicator provides the attack display in terms of range, azimuth, and elevation of the target, and an artificial horizon display of his own altitude.

A fire-control system has many components installed in separate housings. The recommended practice of incorporating as many units as practicable in one housing to minimize interconnecting wiring is not applicable in this system bec[illegible]e of the space limitations in fighter aircraft. Many smaller units facilitate installation engineering. However, the resulting maze of interconnecting wiring magnifies the radio interference problem.

Components of such a system would include: Antenna, radar modulator, receiver-transmitter, servo amplifier, computer, range servo, dimmer control, rocket-setting control, sight head, range indicator, power supply control, power supply, manual range control, gyroscope, electronic control amplifier, gun-laying radar central, two azimuth-elevation indicators, antenna (hand control), blowers, wave guide, and cabling. The main sources of interference in this typical fire-control system would include the antenna, receiver-transmitter, and the modulator.

The antenna unit which includes potential interference generators such as relay

switches, servo motors, and drive motors, necessitates careful consideration in the design stage. In this system's original design, relays were suppressed by the addition of capacitors in parallel with the relay switch. Feed-through capacitors (0.01 µf) were also included in series with the relay input leads. However, circuit arrangement was such as to couple interference back into the filtered lead nullifying the suppression techniques. More effective suppression could have been had with the addition of a resistance-capacitance circuit in parallel with the switch rather than a capacitor; proper routing of leads would have eliminated coupling of interference back into the "clean" leads. Filters were installed on the output leads of the servo motors. However, again, the "clean" leads out of the filter were exposed to interference coupling.

The receiver-transmitter (RF Unit) proved to be an indirect source of interference currents by providing a coupling path for interference from the modulator pulse. The pulse entered the receiver-transmitter through a shielded cable and the interference coupled through the circuits to appear finally on the external wiring enabling the interference to couple into other airplane wiring. Improved shielding or filtering techniques within the receiver-transmitter could have eliminated this coupling path. Case construction of the receiver-transmitter was poor because it utilized a rubber gasket for its pressurized seal. Radiated interference from currents originating in the modulator was leaking from the case; conductive gasket material substituted for the rubber gasket would have eliminated this source of leakage,

The pulse cable between the modulator and the receiver-transmitter proved to be a severe source of radiated interference. A loop probe measured 12,000 microvolts of radiated interference. The cable used was of approved coaxial type with four layers of shielding; however, the connectors or plugs provided poor shielding continuity with the cable, and leakage caused interference to appear on all parts of the coaxial line and to radiate or couple into other airplane wiring. A different type of coaxial line and plugs were used and materially reduced the interference. The coaxial cable replacement had inferior shielding properties (2 layers of shielding); however, good continuity was achieved with the new plugs. Better results could have been obtained if the original cable had been used with an improved type of plug.

The modulator in this system, a source of interference by the nature of the function it has to perform (see Paragraph 3.2.2.3 on Modulators), incorporated widespread shielding of internal wiring. The effectiveness of this shielding was considerably reduced by leaving about four inches of pulse cable unshielded within the case from the plug to the capacitor. This enabled interference to radiate to all susceptible leads and circuits within the modulator and also to be radiated and conducted from the case. The most efficient pick-up of this interference was by the coils of relays and blowers. Subsequently, the interference could be conducted out of the case by the relay and blower, power and control leads. Shielding effectiveness was further reduced by not adequately bonding the shielding conduit and by using throughout the modulator what amounted to "floating" shields. (See Paragraph 1.8.2.2.) As in the receiver-transmitters, the modulator employed a rubber gasket to maintain its pressurized seal which permitted radiation of interference from the case. Shielding braid substituted for the rubber gasket again would reduce this case radiation.

Poor design practices observed throughout the set would include:

(a) "Noisy" and "clean" leads were bundled together, a malpractice which permitted coupling of interference into "clean" leads.

(b) Filter and by-pass capacitors were made ineffective by routing the filtered lead back through the interference field.

(c) Filters were not mounted close to the interference source and the portion of the lead between the source and the filter was left unshielded and free to radiate.

(d) Shielding braid was not grounded.

(e) Ground wires were unnecessarily long, a poor design practice which creates high impedance paths and permits radiation of radio interference from these ground leads.

To illustrate the importance of proper routing, grounding, and shielding techniques, two tables have been prepared which show conducted and radiated interference before and after the following remedial actions were taken:

(a) "Noisy" leads were shielded, or when shielding was impractical, rerouted or filtered.

(b) Rerouting of leads coming out of interference suppression units was accomplished so as to avoid coupling of the original interference back into the "clean" lead.

(c) Filters were mounted as close to the interference source as possible.

(d) All "floating" shields were effectively grounded at both ends and at intermediate points when necessary with leads whose length was no longer than two inches.

(e) Other ground leads, some as long as 10 feet, were eliminated and short leads (2 inches) substituted.

(f) The pulse cable and plugs were improved and properly grounded.

(g) Shielding braid was placed between the mating surfaces of the modulator and the receiver-transmitter.

(h) Leads conducting interference out of the modulator and receiver-transmitter were filtered at a point as close to the pin connector as possible.

The results of the tests demonstrated in Figure 3.3.1.11-A and Figure 3.3.1.11-B were conducted in accordance with Specification MIL-I-6181. The corrective action taken on this fire-control system was not intended to be complete, but merely to demonstrate the appreciable improvements that can be had by proper design considerations.

The interaction of this equipment with other aircraft systems can be demonstrated

RADIATED INTERFERENCE IN MICROVOLTS BEFORE SUPPRESSION TECHNIQUES AND AFTER				
	Before		After	
Freq. in MC	Search	Hand Control	Search	Hand Control
0.16	31	80	8	7
0.18	37	100	12	12
0.24	60	145	12	12
0.28	85	160	13	13
0.32	100	180	11	10
0.6	85	140	10	10
1.0	54	100	9	8
1.4	26	58	3	3
1.8	7	34	3	2
2.8	16	16	2.5	2
3.8	11	12	2	1
5	10	11	4	4
7	11	15	2	2
9	5	5	2	1.5
11	14	14	1.5	1.0
14	14	14	2	2
16	12	15	2	1
18	65	15	2.5	1
20	4	5	3	2
30	100	100	80	30
32	250	380	90	40
36	50	60	30	18
38	450	750	50	18
40	180	250	22	12
45	50	60	45	8
50	100	115	50	22
65	110	120	70	70
75	220	160	30	30
80	60	50	50	50
90	55	45	50	45
100	300	86	18	10
110	55	38	5	5
120	60	58	22	22
140	100	110	24	20
150	80	90	15	10

Fig. 3.3.1.11-A

CONDUCTED INTERFERENCE IN MICROVOLTS
BEFORE SUPPRESSION TECHNIQUES AND AFTER

Freq. in MC	Before AC Line		After AC Line		Before DC Line		After DC Line	
	Search	Hand Control	Search	Hand Control	Search	Hand Control	Search	Hand Control
0.16	40	40	25	25	55	55	2.5	1
0.18	60	60	50	50	70	70	3	3
0.24	55	55	45	45	140	140	1	1
0.28	57	50	45	40	60	58	1	1
0.32	90	92	45	45	75	70	1	1
0.6	28	28	20	18	9	6	1	1
1.0	28	28	7	7	9	9	1	1
1.4	6	6	6	5	10	5.5	1	1
1.8	4	4	3	2	14	25	1	1
2.8	11	11	1.5	1	18	5	1	1
3.8	20	18	1	1	11	8	1	1
5	40	37	2	2	30	22	2	2
7	20	18	4	4	80	60	1	1
9	23	25	2	2	27	28	2	2
11	65	60	2	2	75	72	2	2
14	100	75	4	3	150	120	20	18
16	65	55	6	5	90	60	5	4
18	50	50	7	3	70	55	2	1
20	20	18	2	1	30	30	3	2

Fig. 3.3.1.11-B

by the following installation difficulties which also point out other important design considerations.

In a fighter installation, two troublesome sources of radiated interference were the antenna spin motor and the thyratrons in the antenna drive-motor circuits. The spin-motor incorporated a current-interrupting governor, the contacts of which proved to be an efficient interference-generating device - affecting the liaison and intercommunication equipment. Another motor was tried and while commutator interference was higher than in the original model, the interference pick-up, when installed, was negligible. Apparently, the type of interference from the second motor (commutator) was more easily suppressed than in the original model. Proper original motor design would have eliminated the need for modification (see Paragraph 3.2.1.1 on Rotating Machinery).

The thyratrons fed into a changing load and standing waves were unavoidable on the leads from the thyratrons to the antenna drive motors. Thyratrons, a non-linear device, also produce steep wave fronts and, consequently, are rich in harmonics. External fixes to suppress interference originating from thyratrons would include shielding, bonding, and filtering of leads carrying the interference. Internally

suppressor circuits such as resistance-capacitance and inductance-capacitance combinations should be installed as close to the thyratrons as possible. When equipment is designed that includes the use of thyratrons, consideration should be given to their interference characteristics. Maximum isolation, most efficient shielding, suppressor circuits, careful routing of leads, should all be included in original equipment design.

In another typical installation, "noise" clicks occurred in the intercommunication amplifier during radar scanning. The clicks were a result of the switching action of the relays controlling the antenna-drive motors. Shielding of the drive-motor circuit leads from the antenna to the radar control and from the antenna to the operator's indicator was necessary. The general effect of the shielding was to reduce the maximum interference values about 50 to 90 percent, at which level the clicks became barely perceptible. The most efficient way would be to incorporate a resistance-capacitance circuit across the relay suppressing the clicks at the source, and eliminating the need for heavy and expensive shielding of external wiring.

3.3.1.12 GUN-LAYING RADAR SYSTEM

A gun-laying radar system is a light-weight equipment used for tail protection of bombers. It operates in the microwave region and can perform the function of "searching" and "tracking" at the selection of the operator. It is similar to the fire control system, also discussed in this section, except that gun laying works with a turret while a fire control system operates in conjunction with fixed guns. Space consideration would not be as much of a factor in a bomber installation and would make possible grouping of units and elimination of interconnecting wiring. The principles of good design incorporated in this system would be applicable to the system discussed previously. The system's essentials are the same. Gun-laying radar sets include six major assemblies: The radar central (with a number of sub-assemblies), the RF unit, the antenna, the indicator, the radar junction box and the indicator junction box.

This system is unusual in that interference suppression was included in the original design as contrasted to the equipment discussed previously. On the first developmental model, without consideration for interference suppression, high levels of both conducted and radiated interference were measured. As discussed in other radar systems, the modulator was the main source of interference. In this installation, it generated high-voltage, short pulses supplied through a pulse cable to the RF unit. High levels of conducted noise were measured on the thyratron and rectifier transformer, AC power leads, the plate power lead, and the common (filament and plate) AC power leads.

A systematic study of the developmental model was conducted with the objective of building a second model which would include good design practices pertinent to interference suppression.

The modulator, a potential source of interference in any radar system, was completely redesigned with the U-shape layout changed to a rectangular design with the tubes and transformer in separate sealed compartments filled with insulating oil. This insulating-oil technique did not prove a practical one because of maintenance difficulties. In order to replace the oil after any repair work, it was necessary to

prevent any air bubbles or moisture from getting into the compartment which proved to be a difficult task. Solid copper shielding was used on all wiring coming from the transformer compartment. Ceramic capacitors (0.001 μf) were used from each terminal to ground. The filament and plate transformer were electrostatically shielded. Filters were installed in the filament supply, plate supply, 115-volt AC lines. Tinfoil gaskets were placed under the covers in the transformer and tube compartments. Components and leads were rearranged within the modulator to isolate high pulse currents from power leads and power supply components. As a result of this redesign, the weight of the modulator was increased from 18 to 24 lbs; however, by the use of more compact design, the dimensions remained about the same.

The pulse cable from the modulator to the RF unit is a frequent source of interference due to leakage. This leakage can be caused by inadequate shielding properties of the coaxial cable, or, more commonly, from poor shielding continuity of the plug-coaxial connection. In this equipment, special plugs were designed to minimize this leakage. It was found that by using a cable with its shield composed of four layers of braid, a reduction of the residual interference from peaks of 50 microvolts to less than 10 microvolts could be obtained.

The RF unit incorporated design features to minimize interference generation as follows: The pulse transformer and the magnetron were built into one unit. The magnetron section of this unit was pressure-sealed while the pulse transformer section was oil-filled and sealed. Special leakage-proof connectors were used for the triggering pulse. Filters were added to the blower motor, relay bus, and two in the secondary of the filament transformer.

In the synchronizer, all pulse cables, e.g., gating, timing, ranging, etc., over three inches long were shielded or rerouted to eliminate possible coupling into "clean" leads.

Two capacitor filters were added to the primary power sources in the radar central, one in the DC supply and one in the 400-cycle supply. A 7-ampere filter was added to the 27-volt lead in the radar junction box. One filter, in each of the three regulated power supply leads, 315, 150, 300 volts, is located in the radar central mounting rack.

In the radar junction box, the wires on the terminal board were rearranged to avoid pick-up from adjacent terminals. Harnessing was rearranged and some shielding provided for high-impedance lines. Resistance-capacitance filters were placed across microswitches (see Figure 3.3.1.11-C) with shielded leads to the filters. Two of the switches were redesigned to provide low-impedance paths to ground from the switch terminals when they are not in active use.

Suppression measures were taken on all "noise" motors in the system. Constant armature-current DC motors were used for scanning and spinning; with the use of filters and the grounding of the negative lead, interference was reduced to well within specification limits. Three 0.25 μf filter capacitors were used in each pair of servo motors. See Figure 3.3.1.11-D.

To suppress interference above 15 mc, two small coils of 1 to 2 μh were placed in series with the armature and two 0.001 μf capacitors were placed across the brushes.

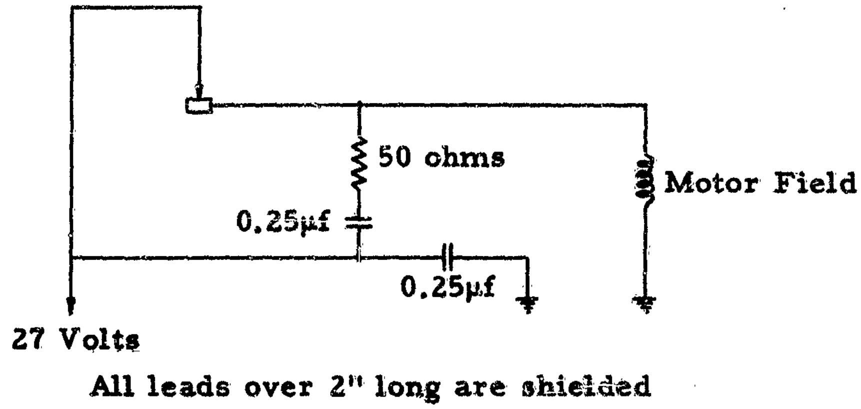


Fig. 3.3.1.11-C Filtering Arrangement for Microswitches

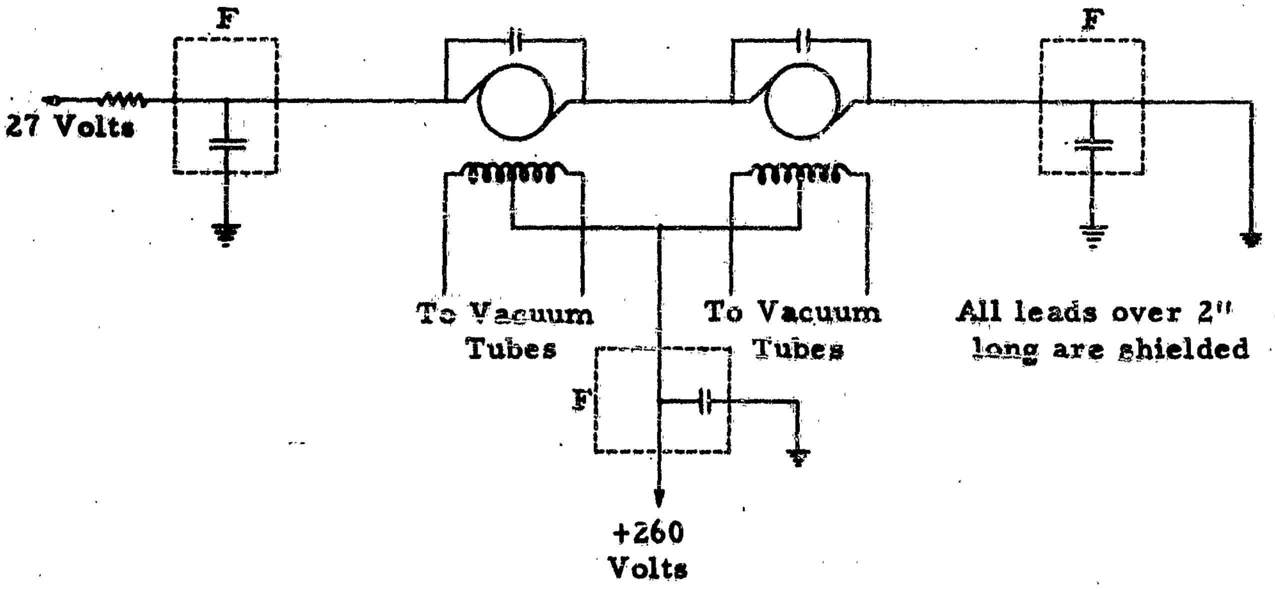


Fig. 3.3.1.11-D Filter for Servo Motors Above 15 MC

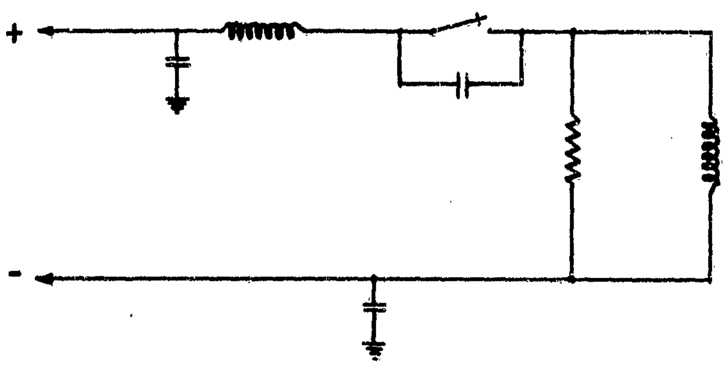


Fig. 3.3.1.11-E Filter for Motor-Control Relay

To save space, each coil was wound around a capacitor and the assembled filter placed within the motor casing. Blower motors were all filtered as close to the motor as possible with adequate grounding provided.

A filtering arrangement was considered for the motor-control relays. See Figure 3.3.1.11-E. While this arrangement was satisfactory in its suppression characteristics, space limitations prevented its use. (See Paragraph 3.2.3.3 on Relays.)

All leads two inches long or longer showing 70 microvolts or more of conducted interference were shielded when possible. Pulse-carrying leads merited special attention to avoid affecting the pulse shape; low-capacity shielded leads were used when necessary. All servo motor leads were also shielded. Non-shielded high impedance leads were by-passed to ground by capacitors.

Careful consideration was given to adequate grounding throughout the system. It was found that anodizing, usually specified for aluminum parts, produces an insulating film which interferes with proper grounding techniques. The following protective coatings are recommended as substitutes for anodizing: chrom-aluminum, zincate process, caustic dip and zinc plate.

The total weight increase, emphasizing internal modifications, amounted to approximately 10 lbs. It is estimated that if these internal preventitive measures had not been taken, the increase in weight for shielding of interconnecting wiring alone would have been in excess of 50 lbs. Furthermore, the interference suppression measures incorporated in this system do not take advantage of all present day approved techniques; e.g., the miniaturization of suppression devices which would have eliminated some of the discussed difficulties encountered in the design.

3.3.2 ELECTRICAL SYSTEMS

Electrical systems are primarily generators of interference transients and fields as a result of the operation of the electrical components of each system. Without the application of any suppression techniques, interference would be conducted and inductively or capacitively coupled to electronic systems throughout the aircraft. In general, an attempt should be made to suppress the interference generated by components at the source. This involves improved mechanical design of the component, adequate shielding, filtering and bonding. Since it is not always feasible or practical to adequately filter all "noisy" components, some interference is conducted through the interconnecting wiring and couples into the electronic systems. Filtering or shielding at points, such as junction boxes, may be sufficient to eliminate these coupling paths. Particular attention should be paid to the routing of interconnecting wires of each system so that "clean" leads are not bundled with "noisy" leads or routed near strong pulse or harmonic generators. Analysis of these electrical systems generally serves to illustrate the type and location of interference-generating components to permit logical references to those paragraphs wherein the components are considered in detail for source suppression. Thus these electrical systems serve to incorporate information available from interference checks on military aircraft as to the types of equipment generally responsible for the production of interference. The following paragraphs describe typical installations to point out some of the general considerations.

3.3.2.1 AC AND DC POWER SYSTEMS

Modern military aircraft are controlled and operated largely through the use of the many electrical and electronic systems installed in the aircraft. A typical medium or heavy bomber would contain the following electronic systems: (1) HF Command, (2) Homing, (3) IFF, (4) Interphone, (5) Liaison, (6) Localizer and Glide Path, (7) Loran, (8) Shoran, (9) Marker Beacon, (10) Navigational Radar, (11) Search Radar, (12) Radio Altimeter, (13) Radio Compass, (14) VHF Command, and (15) Jamming Equipment; in addition, the following electrical systems would generally be installed: (1) Auto-Pilot, (2) Flight Instruments, (3) Cabin Heating and Ventilation, (4) Interior and Exterior Lights, (5) Fire Detector, (6) Fuel Booster Pumps, (7) Fuel Gage, (8) Tachometer, (9) Operational Indicators and Engine Instruments, (10) Propeller Feathering, Unfeathering, Reversing, and RPM Control, (11) Propeller De-Icer, (12) Trim Tab Controls, (13) Windshield Anti-Icer, (14) Wing and Tail Surface De-Icer, and (15) Aileron, Rudder and Elevator Controls. The electrical energy required to power these systems is supplied by common AC and DC power circuits. This provides a means for introducing interference signals from either power system into any electronic system as well as a common path between any of the electrical or electronic systems in the aircraft.

The aircraft ignition system is not interconnected with other electrical systems. An isolated ignition circuit complete with power source is provided for each engine. This precludes the possibility of conductive coupling of interfering signals out of the ignition system during operation. A conductive path for interfering signals into the DC power system exists when induction boosters are used for starting. This power is usually broken by the starting relay after the engines are started. However, there is a strong possibility for inductive or capacitive coupling and radiation from the system wiring. Interference problems peculiar to ignition systems are treated in Paragraph 3.3.2.4.

A typical ignition system is shown in Figure 3.3-C by broken lines to illustrate the compactness of the ignition system and at the same time point out the possible paths for interference signals to gain access to the other aircraft systems.

In general, a typical AC power circuit would consist of the following components: (1) Bus Bars, (2) Voltage and Power Selector Switches, (3) Alternators, (4) AC Voltmeters, (5) Compensating Condenser, (6) Voltage-Adjusting Rheostats, (7) AC Voltage Regulators, (8) Circuit Breakers, (9) Inverters, (10) Junction Boxes, (11) Control Panels, (12) Various Plugs, Receptacles and Interconnecting Cables, and (13) Fluorescent Lights. A typical AC power circuit installation is illustrated in Figure 3.3-C. This diagram shows the approximate location of the various electronic systems and how these systems are interconnected through their power leads in the aircraft. The AC power cables are represented by broken lines (dashed).

A typical DC power circuit would consist of the following components: (1) DC Generators, (2) Batteries, (3) Bus Bars, (4) Generator Ammeters, (5) Battery Ammeters, (6) Voltmeters, (7) Voltage Selector Switch, (8) Circuit Breakers, (9) Internal and External Lights, (10) Voltage Regulators, (11) Relays, (12) Condenser, (13) Connector Boxes, (14) Control Panels, and (15) Various Plugs, Receptacles and Interconnecting Cables. The majority of electrical actuators and servos are energized by the DC source. A number of such devices are shown in Figure 3.3-C. This

diagram illustrates the relative locations of the various electrical systems and how these systems are conductively linked to the electronic systems through their respective power leads. Direct conductive links between the various DC electrical systems generally occur in the junction boxes, circuit breakers, and control panels. Since in many typical installations the aircraft generators are AC-DC, a possible conductive path also exists between the electrical and electronic systems. The AC-DC generators frequently utilize the same coil windings for each external system. A commutator is provided to rectify the induced alternating voltage for the DC system and slip rings are provided for the AC system. Some isolation effect is obtained through the use of separate power supplies for the various systems, i. e., dynamotors, inverters, etc. Bundling of the power cables with other system wiring introduces the possibility of inductive or capacitive coupling of interfering signals from one system to another. The DC power circuits are represented by solid lines.

The functioning of the power systems is not impaired by the presence of interfering signals along the power cables or in the system components. However, a great number of interference difficulties have been traced to the location, routing, shielding, etc. of the power cables. These troubles are primarily caused by other systems coupling interfering signals into the power lines and eventually introducing these signals into the various receivers in the aircraft. The generators, regulators, and relays used in the power systems are possible sources of interference. The design considerations applied to aircraft components for minimum generation of interference are discussed in Paragraph 3.2 of this book.

3.3.2.2 PROPELLER SYSTEMS

Propeller systems are potential and actual sources of radio interference because they contain electrical and electronic equipment and circuits. This is true regardless of whether the propeller is of the electric or the hydraulic type since electrical means are employed to accomplish the following diverse functions in present-day designs of propeller systems:

(a) Propeller Control
- (1) Speed control
  - i. Reciprocating engines
  - ii. Turbo-prop engines
- (2) Synchronization
  - i. Speed synchronization
  - ii. Phase synchronization
- (3) "Beta" operation (primarily in turbo-prop applications)*
- (4) Reverse-pitch operation (primarily in reciprocating engine applications)
- (5) Feathering
  - i. Manual initiated
  - ii. Automatic type
- (6) Remote control of speed settings of governors and synchronizers.

(b) Propeller de-icing

*In "beta" operation, the pilot has direct control, through manual operation of the lower lever, of the propeller blade angle. This type of operation is used for taxiing and other ground handling as well as reverse-pitch operation for ground braking.

(c) Propeller blade-angle (beta-angle) indication.

(d) Electrical lamp indication of propeller operations.

Basically the problem of radio interference prevention on such equipment is the same as for other aircraft equipment and the same principles and practices might be expected to apply. Such equipment, incidental to its basic operation, generates rapidly changing currents and voltages. A certain amount of coupling exists in the aircraft due to the proximity of equipment in the necessarily close confines. This coupling comes from two principal sources:

(a) Use of common power system and battery for both the propeller and the radio and electronic equipment, and

(b) Mutual inductance between propeller wiring and antenna lead-ins of radio equipment.

Two factors seem to distinguish propeller systems from other sources of radio interference in aircraft:

(a) The current values used are very high, particularly on 28-volt DC systems, and normal operation involves continual switching on and off of such currents. From a weight penalty standpoint, the margin of error in the direction of supplying a generous amount of suppression filtering is narrow.

(b) Many important sources of interference in propeller systems produce interference of the click- or pulse-type of low and very low repetition rate. This situation is illustrated in Figure 3.3.2.2-A. The practical result is to create a difficult problem in measurement technique.

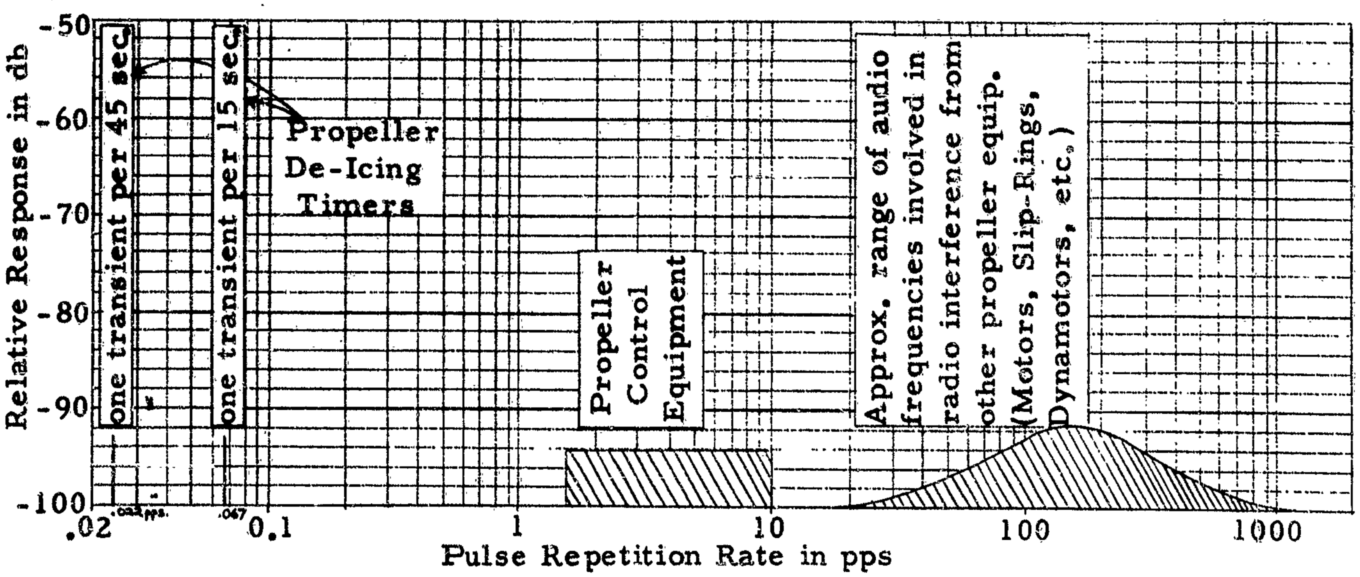


Fig. 3.3.2.2-A Approximate Repetition Rate of Important Sources of Radio Interference of the Click or Pulse Type

The present specified measuring set-up prescribes two standard service receivers. It also involves the use of a cathode-ray oscilloscope as an indicating instrument rather than a D'Arsonval meter used in conventional noise meters. The reason for this approach lies in the wide range of so-called repetition rates which are involved in the types of interference that are produced by propeller circuits as shown in Figure 3.3.2.2-A. Particular note should be paid to the low values of repetition rates involved in the pulse-type "noise" produced by many propeller control systems and the very low values for certain types of de-icing timers. This interference is primarily a series of clicks or pops with "clear" places between.

Sources of interference in propeller systems may be found in (a) pitch change motors, (b) pitch change solenoids, (c) slip rings, (d) governors, (e) synchronizers, (f) de-icing timers, (g) de-icing relays, (h) inverters, and (i) the various switchers and contactors. Many of these items can readily be seen illustrated in Figure 3.3.2.2-B. Such equipment, incidental to its basic operation, generates rapidly changing currents and voltages.

A certain amount of coupling exists in aircraft due to the proximity of equipment in the necessarily close confines. This coupling can come from two principal sources: (a) use of common power system and battery for both the propeller and the radio and electronic equipment, and (b) mutual inductance between propeller wiring and antenna lead-ins. See Figure 3.3.2.2-C.

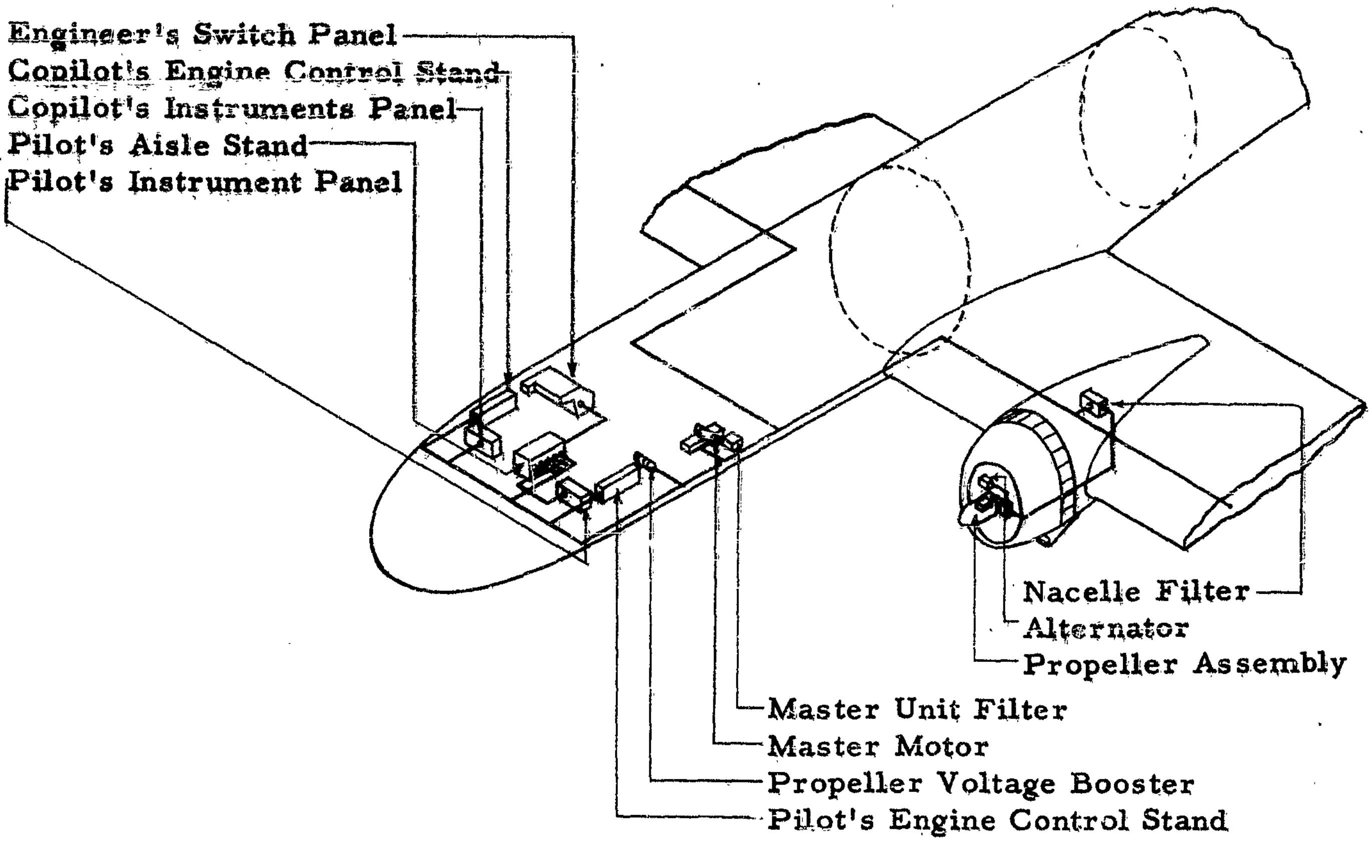


Fig. 3.3.2.2-C Propeller Control Diagram

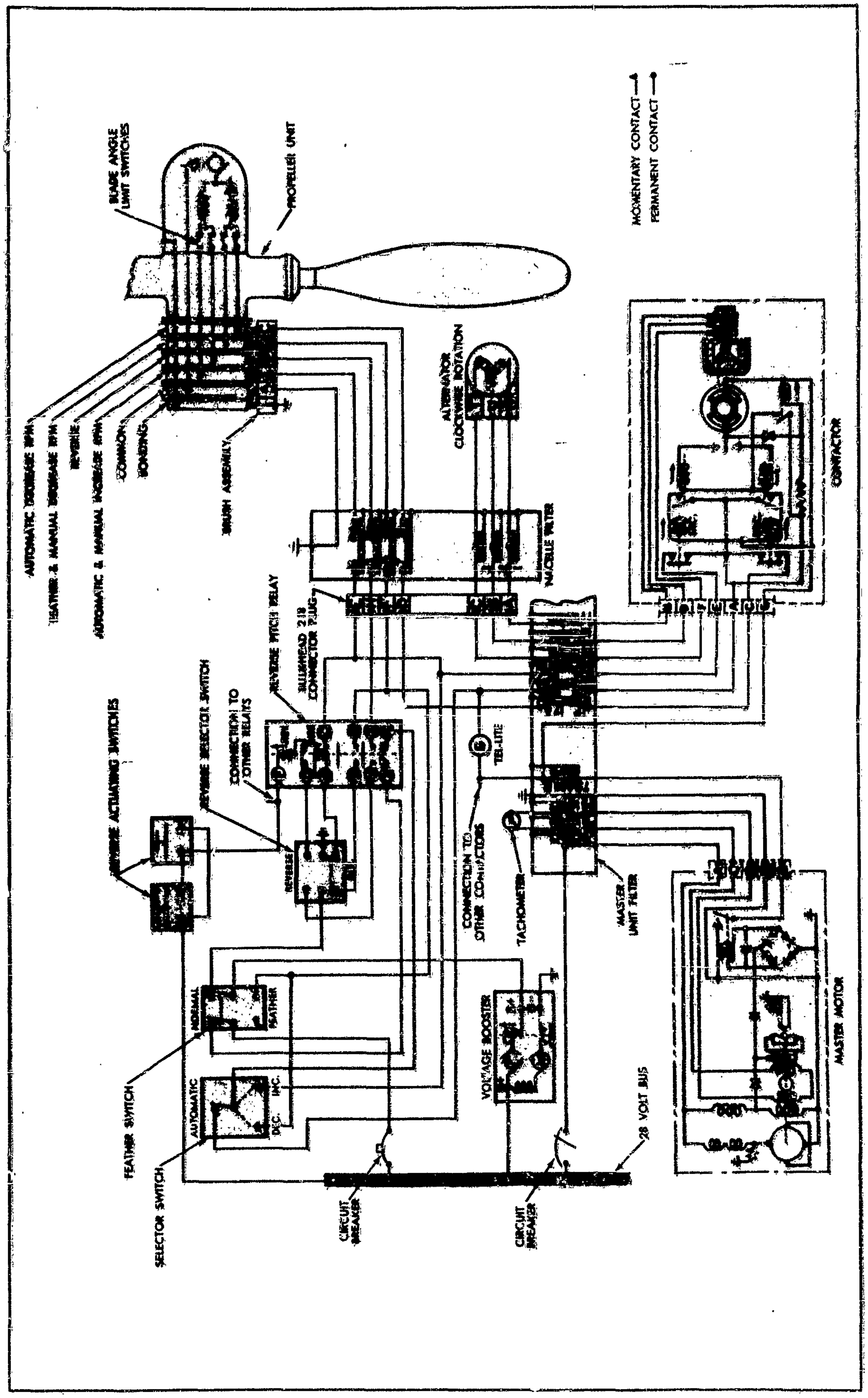


Fig. 3.3.2.2-B Schematic Wiring Diagram of a Representative Propeller System Installed in a Typical Airplane

Experience shows that it is difficult to add radio-interference suppression devices after propeller designs are complete. For that reason, it is far better to design the equipment correctly to begin with. Practical methods of prevention include: (a) use of non-electrical and non-electronic methods of operation in propeller equipment, (b) where electrical equipment is used, arrange the equipment so that circuits involving rapid changes of large currents are placed out in the engine nacelles rather than in the fuselage and particularly not near the radio equipment, (c) in cases where propeller equipment is to be powered from a common electrical system with the radio gear, provide effective filtering in the power leads, and (d) in cases where propeller equipment containing interference sources is to be installed near radio equipment, provide adequate shielding containers for the propeller equipment and filter all leads from the equipment.

The filtering of leads from interference sources in propeller equipment requires the use of the following components: (a) capacitors, (b) inductors, and (c) transient suppressors of the following types: (1) dry-plate rectifiers (magnesium copper sulfide), (2) point rectifiers (germanium), (3) non-linear resistors of the "Globar" or "Thyrite" type, (4) gaseous discharge tubes, (5) vacuum tubes, and (6) resistor-capacitor networks. Capacitors find universal application across lines from interference sources as suppressors; they are particularly effective on DC motors and generators. Addition of an inductor between the capacitor and relay contact makes an effective combination in suppressing relay and contactor interference. Additional inductors in series and capacitors in shunt can be used in the form of radio-interference filters for further suppression. For filtering to be adequate it must be supplemented by the use of shielding containers. Proper shielding involves the use of metallic (not necessarily magnetic) boxes and containers that are free of openings and non-conducting joints. Size and shape should be as required for the propeller equipment. The design techniques for shields and joints are discussed in Paragraphs 1.8.2, 3.1.2, and Appendix XVI of this book.

The designer must be guided by experience in the amount, type, and extent of filtering and shielding to use. The final criterion is that the propeller equipment cause no interference in the radio and electronic gear in the actual aircraft installation.

There are, however, obvious practical disadvantages to waiting until equipment is installed in the aircraft before being able to determine radio-interference performance. The answer to this difficulty has been to set up test specifications setting forth the procedures, measuring equipment and maximum allowable limits to be used in determining the radio-interference performance of electrical equipment for aircraft. Propeller equipment has received its share of attention in this effort.

The thought in setting up a radio-interference specification for propeller equipment has been to provide the manufacturer with a guide as to how good to make his equipment. The expectation is not that equipment meeting such a test will never cause interference when installed; the range of variables involved is too great for that. Rather, equipment, when passed, can be expected to cause interference in less than about 10 percent of actual cases. These latter problems could then be handled individually. Working that close to the margin requires good test equipment and procedures together with maximum allowable limits founded on a sound basis.

In this connection, it may be helpful to review the general basis upon which future plans are founded. This envisages that a radio-interference survey will be made as part of the type test of each new propeller design. This approach would be analogous to the use of vibratory-stress surveys on new designs. Indeed, it is believed that the radio-interference survey can be conducted just before or just after the vibration study. The type test approach was chosen rather than a production inspection method because of conditions which exist in the propeller industry. Production methods in the industry are such that while operational checks are continually made of components of propeller systems, very seldom, if ever, are complete set-ups made during production of systems to check operation prior to installation on the airplane. In many cases new designs of propeller systems consist of only minor changes from previous designs or consist of changes of a mechanical or similar nature not usually affecting the radio-interference performance of the system. It is believed that if a system has been once tested for radio interference and found satisfactory, approval can be granted on various modifications of the system without costly retesting of each modification. A similar procedure is currently used in handling vibratory stress approval on propellers differing in minor detail from a previously tested and approved design. In this connection, attention of contractors is directed toward Specification MIL-P-5449, Amendment 1, requiring the furnishing with the propeller installation model specification of detailed data on the radio-interference suppression provisions.

Two alternative forms for the type test are prescribed: (a) engine tests, and (b) bench tests.

The engine test procedure is to be used for testing all propeller systems and equipment the radio interference from which can reasonably be expected to be influenced by engine rotation, vibration and propeller power loading. Such effects can be expected in systems involving such items as: (a) electrical current-carrying slip rings in the hub or on the blades, (b) engine-mounted propeller governors, and (c) pitch-changing mechanisms of an electrical nature.

The bench test procedure is to be used for testing all propeller systems and equipment the radio interference from which can be expected to be unchanged in magnitude by the presence or absence of engine vibration or propeller power loading. Typical examples of such equipment are: (a) blade-angle indicators involving use of AC selsyns powered by 400 cps inverters, and (b) de-icing system using hub-mounted generators together with a timer.

Manufacturers of propellers and propeller components are cautioned that it is the practice of the services to follow up all complaints of interference and to request modifications, where required, of propeller equipment. Responsibility for performing tests to check compliance of such equipment with applicable radio-interference specification lies with the group having jurisdiction over the equipment. All questions of compliance, test methods, and equipment design should be referred to the group, laboratory, etc., having jurisdiction over the equipment. In the typical case of the USAF, the group having jurisdiction over propeller equipment is the Propeller Laboratory, Aeronautics Division, Headquarters, WADC, Wright-Patterson Air Force Base, Ohio.

3.3.2.3 AUTOMATIC PILOT

An Automatic Pilot is generally an electromechanical-electronic system designed to provide automatic control of airplane surfaces to maintain a predetermined course of flight. The plane is stabilized on longitudinal, lateral, and vertical axes with minimum angular displacement and accelerations in any axis. Manual control of the Automatic Pilot to accomplish dives, climbs, and coordinated banking turns through wide limits is provided.

The components of a typical system include: (1) Flight Control Rate Gyros, (2) Directional Panel, (3) Turn and Pitch Controller, (4) Turn Control Transfer Switch, (5) Three Servo Motors, (6) Multiple Channel Amplifier, (7) Calibrator Unit, (8) Mounting Chassis, (9) Amplifier, (10) Turn Controller, (11) Formation Sticks, (12) Interconnecting Cabling, (13) Flight Control Vertical Gyro, and (14) Gyro Directional Stabilizer.

In a typical installation, in a bombardment type aircraft as shown in block form in Figure 3.3.2.3-A, the control mechanisms which include Turn Control Transfer Switch, the Turn and Pitch Controller, and the Formation Sticks would be readily accessible to both pilot and co-pilot. The Directional Panel is installed in the bombardier's compartment to be used in conjunction with the bomb-sight for flight control during bombing runs. The chassis equipment assembly, which includes vertical and rate gyros, calibrator, and amplifier mounted in a tray, is installed for ease of maintenance. The three servo motors are installed close to the control surfaces and are connected to the amplifier with lengths of electrical cable.

This Automatic Pilot operates on the principle of the balanced bridge. A balanced bridge is provided for each of the three axes: aileron, rudder, and elevator. When any one of these bridges is unbalanced by a signal from any one of the gyros, formation sticks, remote control, or turn and pitch controller, a relay switch in the amplifier activates the appropriate servo to position the control surface to a predetermined or new setting. The process is one of continuous correction with subsequent continuous operation of the relays signalling the servos. The rate of the corrective response is regulated by the calibrator.

The Automatic Pilot, by the nature of the components necessary for its operation, can be a prolific source of radio interference to other systems in an aircraft. The continuous operation of relays and small motors generates interference over a wide range of frequencies, and requires consideration in the original design of such equipment. In the original design of the Automatic Pilot under discussion, interference suppression was not stressed. Examples of poor interference suppression design include:

(a) Mating surfaces were anodized preventing good ground and bond connections.

(b) Bolts, nuts and washers were treated with a corrosion resistant compound which was non-conductive and contributed high impedance paths to ground.

(c) Poor case and internal shielding permitted radiation and coupling of interference into other airplane wiring.

(d) Filters were used with long unshielded leads thus permitting radiation and coupling of noise into other wiring.

(e) "Noisy" leads were bundled together with "clean" leads resulting in interference appearing on most external wiring and making it difficult to apply suppressive measures. Leads coming out of filters were exposed to interference fields which nullified the effect of the filter.

The principle components of the system from which the interference would originate includes the Calibrator, Amplifier, Directional Panel, and Turn and Pitch and Controller.

The calibrator with its numerous relays, e.g., aileron engage, rudder engage, elevator engage, auto recovery, etc., proved to be a prolific source of interference. The action of the relays caused currents with steep wave fronts which resulted in radiation and coupling into other airplane wiring. In the original design, suppression networks were not included in the relay circuits. External filters were inadequately grounded and improperly located to suppress interference adequately. Shielding cups prevented radiation from above; however, interference was free to radiate from beneath the relay. The calibrator case had poor shielding continuity, bonding, and grounding, and permitted radiation to emanate from the case. It was necessary to modify the calibrator internally to prevent it from interfering with other aircraft equipment.

In the amplifier the six servo relay switches were the principle sources of radio interference. It was particularly serious in this component because of the long lengths of cabling to the three servo motors which could cause interference in a considerable amount of airplane wiring in the original design. The only recognition of the interference problem caused by relays and their associated wiring took the form of inadequately shielding each relay internally. Interference radiated from the poorly grounded shield, and transients in the servo wiring affected other airplane wiring. Poor shielding continuity and grounding of the amplifier case also permitted radiated interference to affect other circuits.

In the Directional Panel, interference from the DC Arm Lock Motor Commutator appeared on external interconnecting wiring and was free to radiate or couple into other airplane wiring. In the original design, the motor was shielded but the leads were not shielded or filtered and conducted interference appeared on external wiring.

The Turn and Pitch Controller has three direct current motors used for rudder, aileron and elevator centering. In the original installation, these motors were individually shielded but no internal suppression measures were taken to prevent the commutator interference from being conducted through the shield by the motor leads. These unshielded leads were bundled in with other wiring and decreased the effectiveness of external filters of DC power leads. Coated surfaces used in mounting the motors did not provide adequate bonding and reduced the shielding effectiveness of the cases.

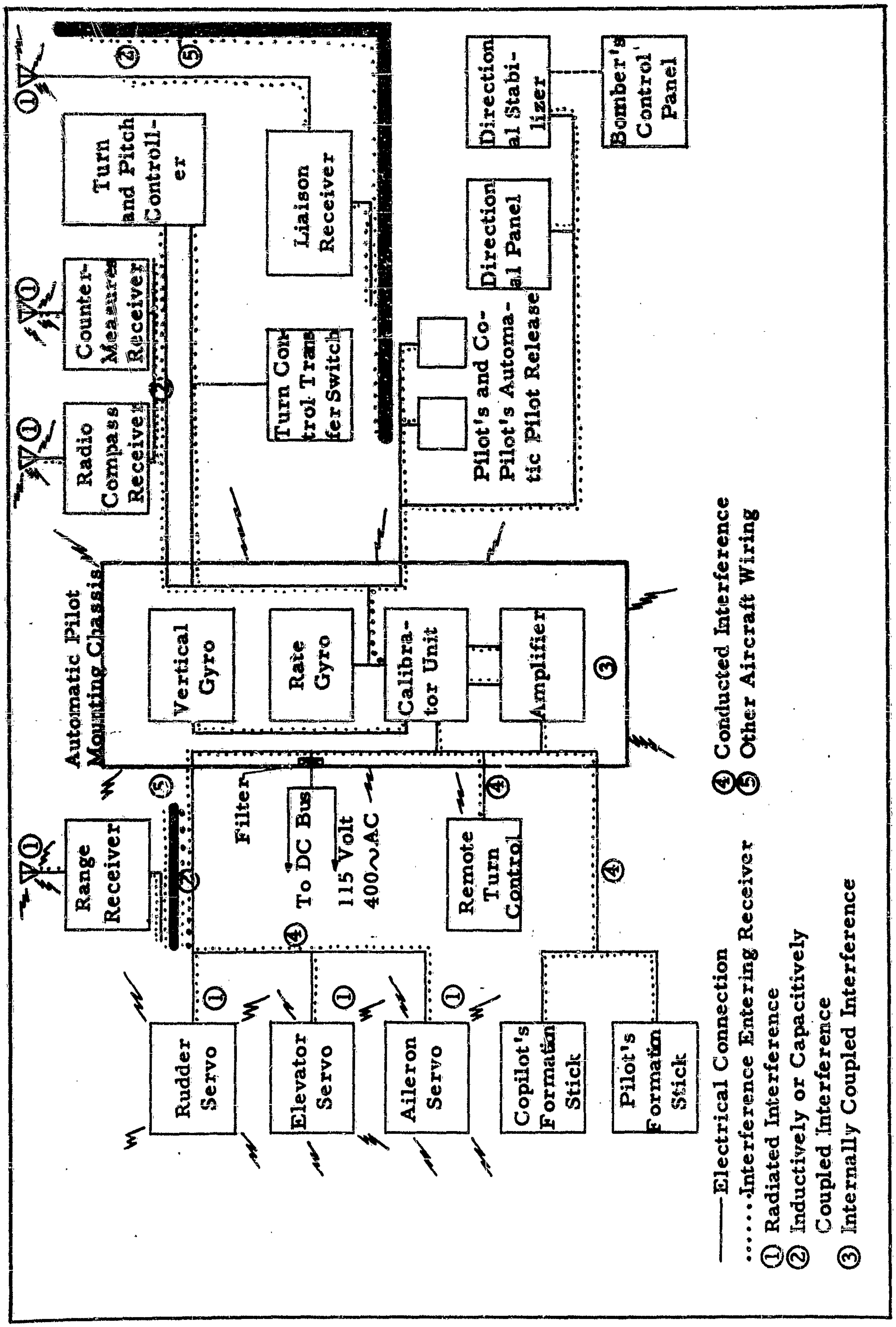


Fig. 3.3.2.3-A Paths of Interference Signals in a Typical Automatic Pilot System

As a result of lack of adequate suppressive measures used in the design stages, installation difficulties were encountered with the operation of the Automatic Pilot in a bomber installation because interference was found on all bands of the Liaison and one of the Countermeasures Receivers. The relays caused a rapid clicking noise in both receivers with peaks measured in the range from 3 to 21 megacycles. Since an internal modification of the Automatic Pilot System was the only practicable way to eliminate the interference, no installation "fix" was attempted. In a later model of the same type of aircraft, the interference from the relays was reduced to a low level by modifications in the Automatic Pilot; however, commutator noise, from the Gyro Torque motor in the Directional Stabilizer and the Arm Lock Motor located in the Directional Stabilizer appeared in the same receivers. A "fix" was attempted by mounting a filter and a 4 microfarad capacitor near the directional stabilizer. The filter was connected in series with the 28 volt DC power lead to the gyro torque motor. The capacitor was connected from the torque motor side of the filter to ground. This "fix" proved to be successful; however, a similar configuration tried on the Arm Lock Motor did not remove the interference from the Countermeasure receiver. The same receivers being affected in both examples of Automatic Pilot interference suggests that improvement may have been possible in the installation mock-up to eliminate the coupling paths for the interference to the receivers. However, this does not remove the premise that inadequate source suppression measures were taken in the original design of the Automatic Pilot.

In another bomber installation, the Automatic Pilot interfered with the Radio Compass, Range and Liaison receivers. The interference was of three types:

(a) The continual or repeated pecking noise from the contacts of the six servo relays. This same type of interference was evident also when the Automatic Pilot was disengaged, but the centering motors in the control panel were being actuated. A "fix" was attempted by using a filter on the load side of each of six servo relays but this only reduced the noise and did not eliminate it. Complete suppression would necessitate modification of the Automatic Pilot.

(b) "Clicks" and "pops" appeared on all three of the affected receivers when the Turn Control was moved in and out of the detent position. Part of the interference was due to the make and break action of the switch contacts at the detent position during the make and break of the Arm Lock Relay coil circuit. This interference was particularly prominent on all except the low frequency band of the Liaison receiver with peaks in the range of 6 to 9 megacycles. A "fix" was attempted by using a 1/2 microfarad metallized paper condenser across the switch contacts. This action effectively suppressed this part of the interference. The remainder of the interference was caused by the Arm Lock Relay Contacts actuating the Arm Lock Motor. These clicks and pops were the sharpest and most annoying of the Automatic Pilot interference and most serious on the Radio Compass and Range Receivers. To suppress this part of the interference, a filter was placed on the load side of each of the three contacts that actuate the Arm Lock Motor. A filter was also placed in the 28 volt power lead to the relay.

(c) The occasional or intermittent "clicks" and "pops" due to the Engaging,

Anti-engaging, Transfer, and Rudder Engage Relays in the Calibrator Unit appeared on all three receivers. Suppression measures were not taken on the Automatic Pilot but an attempt was made to eliminate coupling paths and increase the rejection characteristics of the receivers. The radio compass antenna lead-in was double shielded and a 1/2 microfarad condenser, 400 volts, DC rating was connected on the 400 cycle power lead of the Radio Compass to ground. Bonding of the Liaison and Range receivers was improved and leads rerouted when possible. This attenuated the interference, but did not eliminate it.

External filtering in this installation was not satisfactory because of the lack of circuit isolation. Internal cabling was laced in large bundles with interference coupling throughout the equipment. To be effective, filters would have to be applied virtually to each individual wire of the many wires coming out of the units.

Extensive modification of the Automatic Pilot was necessary in order to insure interference-free installations in aircraft. The following interference-free design techniques were applied:

(a) Good bonding and grounding surfaces were insured by changing the anodizing protective process to cadmium plate over copper plate over zinc plate on the aluminum.

(b) Non-conductive protective coating was removed from bolts and nuts to lower the impedance paths to ground as discussed in Paragraph 3.1.2.7 of this book.

(c) Both internal and case shielding were improved by new design as well as by the previous improvement of contact surfaces as discussed in Paragraph 3.1.2.

(d) Filter ground leads were shortened and the filters were relocated close to the "noise" source with a shielded lead into the filter as discussed in Paragraph 3.1.1.

(e) The internal circuitry was improved which resulted in better isolation of interference sources and interference conducting leads.

(f) Calibrator relay contact surfaces were improved and the leads into the filters were shielded thus preventing the relay interference from coupling into other wiring.

(g) Case shielding continuity, bonding, and grounds were improved to prevent case radiation.

(h) Suppressor networks were installed inside the amplifier and connected to the servo relays. These networks were connected to the signal leads of each of six relays as close to the relay shielded case as possible and a capacitor was connected across the relay contact points to reduce the effect of transients as shown schematically in Figure 3.3.2.3-B. The relay shield was improved as well as overall case design.

(i) The Arm Lock Motor in the Directional Panel was internally filtered with a shielded lead into the filter. This effectively eliminated commutator interference by confining it to its internal shield.

(j) In the Turn Pitch Controller, filters for the three DC motors were more advantageously located, provided with better contacts, and connected to ground by shorter leads. Leads which conducted interference were shielded prior to entering the filters. Rerouting was accomplished to protect the filtered lead from exposure to an interference field. Shielding effectiveness of the case was improved by changing the anodizing to a plating process.

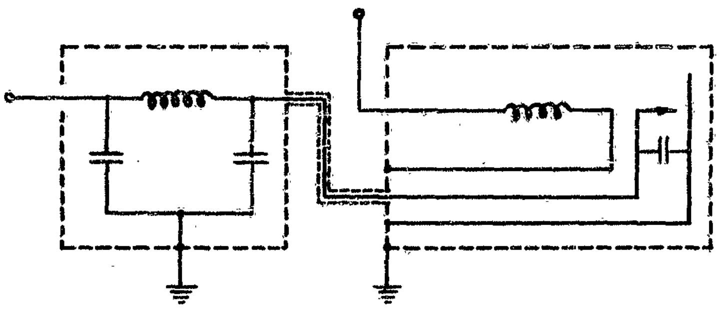


Fig. 3.3.2.3-B Schematic Diagram of Suppressor Network

Up to this point, the interference generating characteristics of the Automatic Pilot have been stressed; now the Automatic Pilot's susceptibility to interference from other systems in the aircraft will be considered.

Interference which entered the Automatic Pilot System in one particular installation caused erratic operation of the control surfaces, known as "jitter". Investigation showed that the signal leads to the servo units picked up interference and conducted it to sensitive circuits in the amplifier at which point the circuits were unable to distinguish the interference signals from the desired ones. By installing a filter in each of the six signal leads, as close as possible to the amplifier, this interference was effectively eliminated. Routing the servo motor signal leads away from interference fields may possibly eliminate the need for these filters; however, owing to the long lengths of cable necessary, filters may be the simplest way to eliminate this as an installation problem. Interference from other systems has not been a major problem in the operation of the Automatic Pilot in this installation.

However, interference from other systems in the aircraft has been a major problem in another type of Automatic Pilot System and a brief discussion of this is included because it still constitutes a problem in recent Automatic Pilot installations.

The principle of operation of the two Automatic Pilots is basically the same. However, instead of a potentiometer type of signal pick-up, rate gyros for the rate signals, and relay pulsing type servo, this equipment uses a synchro type signal pick-up, resistance-capacitance type of rate circuit, and an amplidyne servo system. This system is sensitive to interference on both the AC and DC power systems. In a typical medium bomber installation, oscillation or "jitter" of the control

surfaces was caused by commutator ripple in the DC power supply and low amplitude modulation of the 400 cycle wave in the AC supply.

Two control surface hydraulic booster pump motors produced commutator ripple on the DC system. When these motors operate simultaneously their ripple voltages combine to form a beat effect which caused trouble. This interference eventually coupled into the sensitive circuits of the Automatic Pilot. With low batteries, or with batteries disconnected from the system, the "jitter" is exaggerated but when the batteries are in good condition they tend to minimize the effect of the commutator ripple. The "fix" recommended for this installation was to install a 500 microfarad, or larger, capacitor from the DC bus to ground near the booster pump motor terminals. As much isolation as was possible was provided between sensitive Automatic Pilot wiring and DC power wiring.

In planning the design of future systems and installations of this kind, early consideration should be given to the isolation of sensitive circuits and wiring from the effects of ripple on the DC power supply because this has been an important source of interference to electronic systems in various aircraft. Ideally, the interference characteristics of motors, such as the booster pump motor, should be suppressed in the design stages; however, the rejection characteristics of a receiver to this type of interference should be equally emphasized in its design stages.

The inverter, in this installation, was putting out a low frequency, low amplitude modulation of the 400 cycle wave in the AC power supply. This modulation was aggravated, by the AC ignition system used, to the point where the Automatic Pilot could not distinguish it from the normal sensing signals. The recommended corrective action in this instance was to provide separate sources of AC supply for the ignition system and the Automatic Pilot. To illustrate the importance of a "clean" AC power supply for this equipment, an Air Force specification required that an AC source be provided with no modulation between frequency limits 0.1 to 60 cycles per second and a total harmonic content of not more than 5% of the fundamental. The original design of this equipment should include considerations that make it compatible with the operation of the other systems in the aircraft. With the increasing use of AC power supply in aircraft, this Automatic Pilot system should be able to operate as satisfactorily as other equipment on the same AC supply.

In the design stages of any equipment, recognized interference sources should be properly isolated by means of shielding, filtering, bonding, or other effective means as necessary considering weight, space and materials as discussed in Paragraph 3.1. Study of the original design of this Automatic Pilot equipment revealed that the main reliance was on shielding. The techniques used, however, were not effective in this case. While it is recognized that shielding is frequently necessary to suppress interference from ignition systems, modulators, etc., it is one of the more expensive methods and may result in greater weight than necessary if a combination of techniques are used. For example, a simple R-C network, properly designed and connected across the contacts of relays may have reduced the transients sufficiently to eliminate the need for shielding of each individual relay. (See Paragraph 3.2.3.3.) Proper routing of leads that may conduct interference would still further reduce the need for this extensive shielding. Small motors with interference suppressed by means of techniques discussed in Paragraph 3.2.1.1 and proper isolation would obviate the need for extra heavy shielding.

To overcome the lack of foresight in the original design of this equipment with regard to interference-free operation, good engineering practice was employed in locating interference suppression devices as close to the source of the interference as was possible. The "fixes" used showed a good understanding of and an application of the principles which must be used in combating interference. The result was a considerable improvement in the overall design which reduced the generation, conduction, and radiation of the interference.

3.3.2.4 SPARK-TYPE IGNITION SYSTEM

Ignition systems used in present day reciprocating aircraft engines are designed to produce an electric-spark to ignite the compressed fuel inside the cylinder. The ignition system must deliver a spark at the proper instant in each cylinder to give smooth operation. Reliability and efficiency are prime design considerations and extreme environmental conditions must be provided for in the design of aircraft ignition systems. The system must operate at temperatures of -40° to -60° F. as a lower limit. At the other extreme, however, the magneto is exposed to 250° F., the cable to 350° F., and the plugs to 500° F. Since the system is mounted on the engine, all parts are subject to severe vibration at frequencies below 200 cps, with forces as high as 75 g's.

The reduction of dielectric strength and ionization potential which accompanies increase in altitude is very damaging. Corona discharg- forms ozone and nitric oxides. The oxides combine with moisture to make nitric acid, which corrodes all metal parts. Both corona leakage and weakened dielectric combat attempts to increase ignition voltage and thereby gain more effective ignition. To avoid the detrimental effects of altitude, the ignition system has been pressurized in some designs and in others it has been filled with a sealing compound. These aging, breaking, or other damaging effects due to environmental conditions generally tend to increase the interference problem created by the ignition system.

The aircraft ignition system includes the following component parts: (1) magnetos and distributors; in some systems these are integral (two per engine), (2) harness assembly and spark plug leads, (3) external transformer coils, used only with low tension ignition, (4) spark plugs (two per cylinder), (5) starting booster coil or induction vibrator, (6) ignition and magneto grounding switch, and (7) flexible metal conduit used to cover the wiring between the engine ignition system, the ignition switch, and the starting assemblies. A front view of a typical ignition system installed on a radial aircraft engine is shown in Figure 3.3.2.4-A.

While radio interference signals entering the ignition system will not adversely affect the system itself, the components of the ignition system are extremely strong interference sources and could cause serious interference in the aircraft electronic systems if proper suppression techniques are not observed.

In general, spark-type ignition systems, if allowed to radiate, can be listed high among the major sources of radio interference. This is true because of the deliberate generation of high frequency transient currents in the ignition circuit as a necessary function in aircraft engine operation.

Whenever an electrical circuit is opened or closed, such as occurs with a

distributor, there is a transient or variable-current state immediately following the making or breaking of the circuit, during which current is either rising or falling. These "steep wavefront transients" are not of the simple sinusoidal type, but have been shown to consist of a fundamental wave of relatively low frequency, upon which are superimposed a multiplicity of higher frequency components or transients.

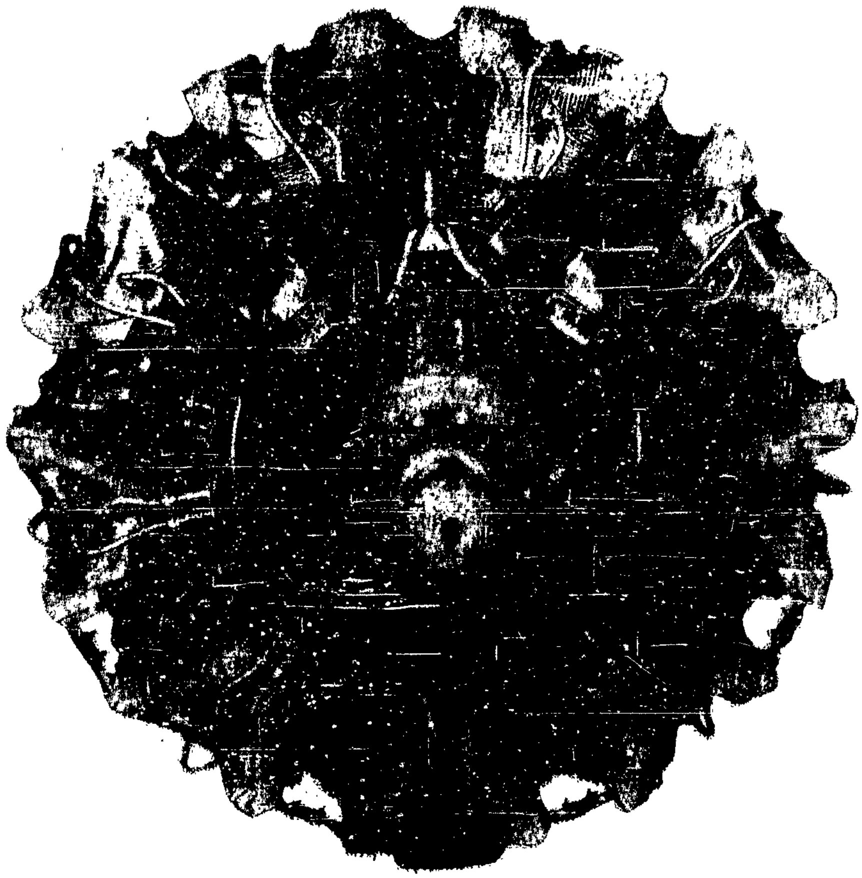


Fig. 3.3.2.4-A Typical Spark-Type Aircraft Ignition System

The components of an ignition system must, therefore, be regarded as generators of periodic short duration waves containing not only low frequency components, but high frequency components extending across the useful radio-frequency spectrum.

These high frequency oscillations, if not suppressed through shielding, are radiated by the various ignition system components which act as antennas. The radiation results in uncontrolled frequency waves in the radio frequency spectrum above 10 mc and is especially high in annoyance value depending upon the characteristics of the transient and the sensitivity, frequency response, and location of the receiver.

Although, as pointed out in earlier paragraphs, noise suppression may generally be accomplished by filtering or shielding at the source or the receiver, or both, such is not generally the case with respect to interfering waves generated by an engine ignition system. With the various aircraft antennas subject to direct uncontrolled frequency radiations of the engine ignition system, any effort to filter or shield at the radio receiver will result not only in the suppression of interfering

waves but also of the desired intelligence signal radiated by a controlled frequency radio transmitting station. The possibility of suppression by filtering at the source is equally remote. Any filter capable of eliminating all of the high frequency components of an ignition transient would also be capable of removing a large part of the ignition wave itself, thereby impairing the efficiency of the entire engine ignition system. Some recent success, however, has been found through use of resistor-type spark plugs.

It follows, therefore, that complete shielding of the aircraft ignition system is the only practicable method to prevent radio frequency energy of appreciable magnitude from radiating into space and resulting in aircraft radio interference.

The existence of this condition is fairly well accepted, although not necessarily appreciated by the industry. As a result radio interference emanating from the aircraft ignition system still is troublesome. The detectable interference now generally results from (a) poor shielding joints, (b) poor flexible shielding conduit, and (c) insufficient shield wall thickness and/or improper material.

In the transmission of high frequency transient currents in a coaxial system, such as is the case in the modern ignition system, the current tends to follow the path of least impedance. As a result, the current will flow on the outside surface of an ignition cable conductor and on the inside surface of the shielding. Any poor junction in the shielding will provide an opening to the shielding exterior thereby permitting the current to flow on the outside surface of the shielding where it can radiate. No bonds are required with a perfect shield.

The problem of ignition noise largely reduces itself to one of joint or flange design and frequent bonding of the shielding to the airframe. Major emphasis should be placed upon proper initial joint design for little can be done in the field to make satisfactory corrections of a permanent nature. Obviously, the number of joints should be kept to a minimum. The most satisfactory joining of mating surfaces is by welding, brazing, or soldering. Even a good solder joint will exhibit an appreciable contact resistance and is never as good a conductor as a brazed or welded joint. However, welding and brazing cannot generally be used because of the difficulties of field service and because the heat required causes damage to insulated conductors.

Radio interference signals may enter or leave the ignition systems over any of the paths shown in Figure 3.3.2.4-B.

Radio interference caused by engine ignition systems can be summed up as follows:

(a) Interference due to radiation leaking past poorly mated joints located in any portion of the ignition system,

(b) Interference from damaged portions of the ignition harness assembly, and

(c) Interference from loose nuts or other fastening devices which are part of the ignition assembly.

The magneto is a form of high-frequency generator and consequently all joints

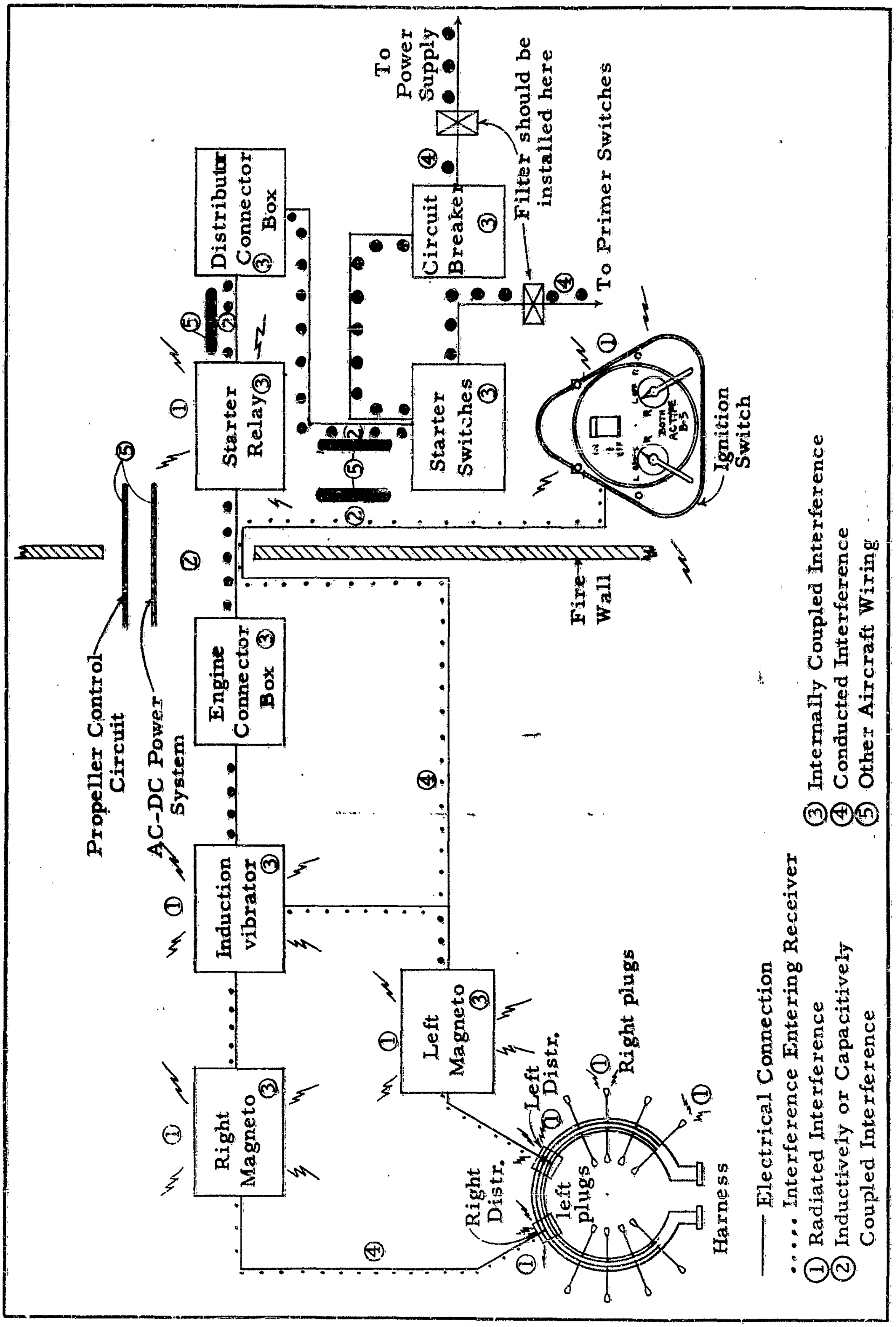


Fig. 3.3.2.4-B Paths of Interference Signals in a Typical Aircraft Ignition System

and covers in the magneto are potential interference sources. Figure 3.3.2.4-C illustrates several types of joints ordinarily found to be a source of interference in magnetos. In order to keep the radio interference energy within the megneto shield it is necessary to make all joints low impedance paths, as indicated above, and discussed in Paragraph 3.1.2.

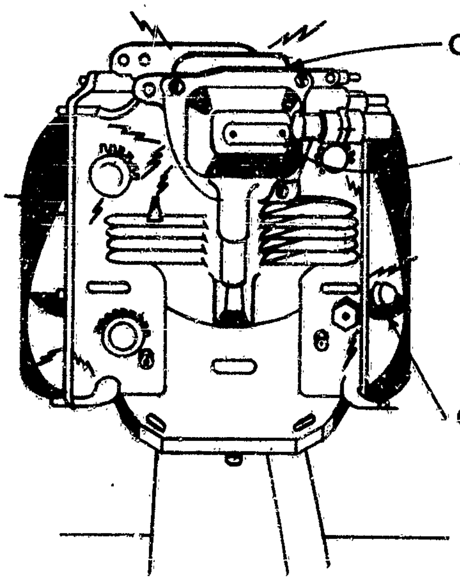


Fig. 3.3.2.4-C Typical Sources of Interference in Magnetos

The distributor comprises a distributor rotor and terminals connecting individual spark plugs. It may be considered as a rotating mechanical switch which transfers the electrical energy from the high tension coil of the magneto to the spark plugs. The distributor finger or rotor does not touch the terminals or electrodes but passes over them with close clearance. Since the high voltage produced by the magneto must jump the air gap between the rotor finger and the terminal of the distributor in addition to the spark plug gap, the distributor can be compared to a spark transmitter. Paragraphs 1.3.2.4 and 3.2.3.1 give a detailed discussion of spark gap interference. Again it is a problem of keeping the joints electrically tight at high frequencies in much the same manner as described for the magneto. Types of troublesome distributor joints are shown in Figure 3.3.2.4-D.

The harness assembly is again a problem of keeping radio interference energy inside the harness or shield. If any portion of the harness assembly is cracked or broken and any connections not properly tightened interference signals will radiate from the harness assembly. Typical potential sources of noise leakage from the harness section are shown in Figure 3.3.2.4-E.

Spark plugs are normally well shielded and seldom present an interference problem, unless they are damaged or not operating properly.

It has been demonstrated by design experience that proper radio interference control of ignition systems is obtainable by using carefully designed shielding assemblies and interference filters. A well-shielded ignition system is well protected

against fire hazard, shock hazard, heat, lightning, weather, and vibration. However, shielding adds to the cost and weight of an ignition system and increases spark plug electrode erosion, thus requiring careful application of suppression techniques in the original design. Shielding and filtering techniques are treated in Paragraphs 3.1.2 and 3.1.1.

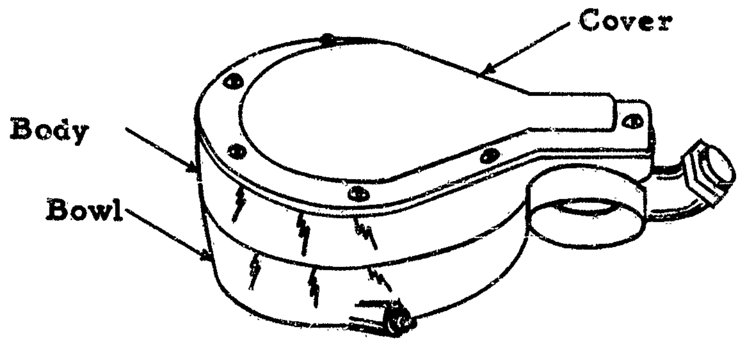


Fig. 3.3.2.4-D Typical Sources of Interference in Distributors

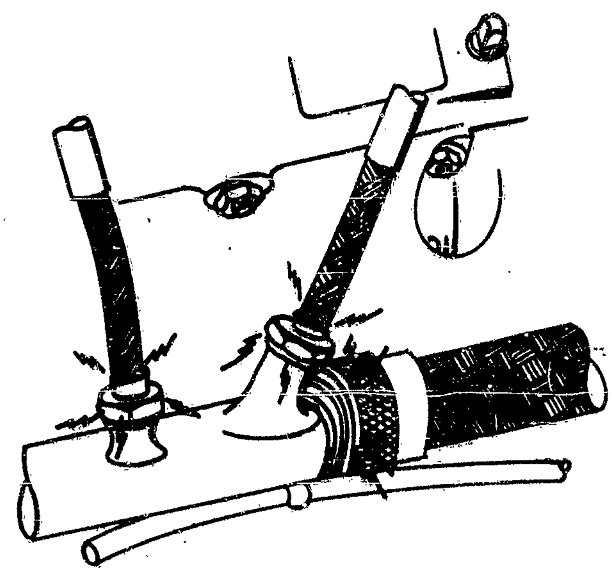


Fig. 3 3.2.4-E Interference Leaks in Harness Assembly

3.4 DESIGN CONSIDERATIONS FOR AIRCRAFT RECEIVERS

To minimize the effects of interference, internal shielding of the individual radio frequency sections is essential. The first radio frequency section of any receiver is the most sensitive and any design feature causing a reduction in interference at this point has a healthy effect upon succeeding stages.

At least 90 percent of all interference enters a receiver through the input circuits of the first radio frequency amplifier stage. Interference conducted into the

receiver by means of the power input cable may enter the first radio frequency stage through the filament circuit as well as through inductive coupling to the internal antenna circuits. The rest of the interference enters this stage through the external antenna circuits, (antenna, antenna lead-in) by means of inductive coupling with the interference carrying wires. The first radio frequency stage must be designed very carefully. Noted improvement is had by adding low-pass filters in series with the first radio frequency filament leads. Lower susceptibility will result from improved shielding design of the internal antenna input circuits. Shielding antenna leads inside the set, and shielding input coils and gang condensers will reduce the coupling to interference circuits, where the shielding is made as continuous as possible. Continuous shielding is obtained quite well by spacing the screws holding down the shield about one inch apart. However, the use of multiple contact serrated springs is much more effective for continuous shielding, as noted under Case Shielding, Paragraph 3.4.1.5.

If the radio-frequency section could be isolated from the receiver and located in a completely interference-free region containing the antenna lead-in, the installation interference level could probably be reduced about 85 percent. The radio frequency and mixer stages could be designed in strips, together with the tuning elements. These could be furnished by the radio manufacturer in standard strips and located by the airplane manufacturer directly adjacent to the receiver antenna lead-in. The design of radio frequency and mixer sections would be completed by enclosing the strips in a shielded case, riveted directly to the metal frame. This method would provide a certain degree of flexibility in receiver design, allowing the airframe manufacturer freedom to utilize the best installation procedures. The power supply, audio and IF sections would be packaged in a smaller case by the radio manufacturer. This undoubtedly requires a certain amount of development work before the procedure could be put into practice, but the work could proceed in conjunction with any installation program initiated to solve space and interference problems. In the design of internal shielding for the reduction of internal coupling, the use of a "one-point ground system" has been found a very effective measure. This means that one point of the internal shield is designated as ground point (see Paragraph 1.8.2.1), and this point is used to terminate all wiring of the shielded stages that need to be grounded. Thus the primary (external) shield functions only as a wall to protect the sensitive circuits inside from the interfering fields outside and does not have to carry any circuit currents. Even in the internal shield, the length of the paths of the circuit currents is kept to the absolute minimum. Thus, the single ground point avoids the impedance which may exist at high frequencies between various ground connections.

An effective way of constructing an internal shield to separate two stages is the use of a copper shield with a circular hole through which a tight-fitting metal tube may be mounted. If the tube is of the type having a grid cap at one end and all other connections at the other end, its metal envelope effectively closes the hole in the shield, and there is complete separation of the grid or input and the plate or output circuits of that stage. Such an arrangement, used for the separation of the radio frequency amplifier from the mixer stage in a superheterodyne receiver, is shown schematically in Figure 3.4.

This method of tube mounting affords excellent isolation of stages to prevent stray coupling. The partition shield may serve as a common ground point for cathode

and one side of the filament, and for plate and grid circuit returns. The tuning condensers may be insulated from the primary shield box, and copper straps used to connect the frame directly to the common ground on the partition shield. The control shaft of the condenser may be brought out of the primary shield box through a close-fitting brass sleeve, which will be soldered to the walls of the box. Proper proportioning of the shaft hole and the length of the sleeve will make the brass sleeve act as an attenuating wave guide, as explained in Paragraph 3.1.2.2.

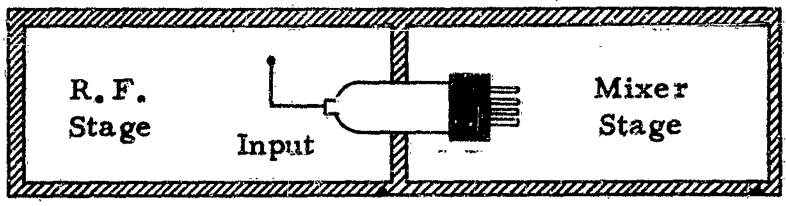


Fig. 3.4 Mounting of Vacuum Tube Through Internal Shield

Interference problems in existing aircraft receivers have been solved in varying degrees by means of fixes. However, the best approach is to design, produce, and install receivers in an aircraft to function with minimum susceptibility by means of appropriate routing of cables, shielding, filtering networks, and by adding circuits to the receiver to materially lower the amplitude of the offending interference voltages.

Aircraft receivers, in general, are similar to other types of radio receivers. Operating conditions, however, require that certain characteristics are to be stressed more than others for minimum receiver susceptibility. The major characteristics are: (a) sensitivity, (b) selectivity, (c) audio-range, (d) bandwidth, (e) shock mounting, and (f) susceptibility.

A high degree of sensitivity is a necessity due to either a short antenna or long range pick-up. Consequently, normally insignificant interfering signals picked up and amplified by the receiver merit serious consideration.

Excellent selectivity is highly desirable for the purpose of eliminating interfering signals. Modern aircraft almost always use a superheterodyne receiver and quite frequently a low intermediate frequency is employed to aid the selectivity characteristic.

A limited audio-frequency range, sufficient for communication purposes, decidedly contributes to interference reduction in a communications receiver output. A range of 300-3000 cycles per second is entirely adequate, and the narrow bandwidth effectively rejects interference impulses over a proportionately greater area of the carrier.

A bandwidth consistent with the minimum possible value for received signals is highly desirable. This makes possible a high signal-to-interference ratio since interference energy is normally distributed over a considerable frequency band. The disturbing effect of background interference can be lowered by volume compression

of transmitted signals and by a reduction of receiver sensitivity during silent periods. (See Special Circuits, Paragraph 3.1.4.)

The Naval Development Research Center conducted experiments at Harvard University showing that the lowest reliable signal-to-interference ratio for intelligible reception of radio signals is 4:1, indicating that coupled interference must be kept very low.

Shock-mounting is another important consideration for receiving equipment to prevent rapid deterioration of operating characteristics as well as actual damage to the equipment. Circuit components must be carefully designed and mounted to withstand shock and vibration. The use of non-microphonic tubes, special capacitors, and proper bonding techniques must be considered to prevent loss of intelligibility due to changed characteristics.

Susceptibility of a receiver to interference is a measure of undesirable response of the receiver to interfering voltages at all paths of entry. It is expressed in terms of "coupling factor" or "susceptibility ratio". Coupling factor may be defined as the ratio of the antenna input voltage to the voltage input required at the various coupling paths to produce the same receiver output and is an index of the receiver's ability to reject conducted interference. Susceptibility ratio, which is sometimes used, is the inverse of this and may be measured as its reciprocal.

3.4.1 PATHS OF ENTRY

The major paths through which interference energy can enter a receiver are: (a) antenna lead-in, (b) power leads, (c) control leads, (d) audio output leads, and (e) receiver case.

Tests have been conducted on representative communication receivers to determine the attenuation on their various paths of entry. The receivers were peaked to maximum sensitivity and maintained in this condition throughout the tests. The sensitivities were recorded so that they could be compared with the voltage input through these paths of entry required to bring the receiver output up to a standard ten milliwatt level. The output of a standard signal generator was applied to the receiver antenna and adjusted for a ten milliwatt receiver output. Then the generator output was applied to a path of possible interference entry and adjusted to give the same receiver output. The ratio of the generator output required for the interference path to the antenna input required for the same receiver output level represented the attenuation of the entry path. The values shown in the following table list the average attenuation in decibels below the antenna input for each of the tested paths of the receivers.

The "A" receiver is a single-band set with a separate power supply. The controls are remote, but the interference enters in the same way as with the other receivers, through the shield, the audio leads, power leads, and a radio frequency sensitivity control. Attenuation is much greater in this receiver due to tighter external shielding, and to shielded radio frequency coils and gang condenser.

	Average Attenuation in db Below Antenna Input		
Paths of Entry	Receiver A	Receiver B	Receiver C
DC Power Input	65	45	40
Sensitivity Control Cable	75	--	30
Audio Output Cable	100	45	45
DC High Voltage Cable	--	30	40
Miscellaneous Control Cables	100	--	50

The "B" receiver has six bands in a single case, together with dynamotor and all controls. The only external wiring are the antenna, power input, and audio cables. Receiver "C" is a single band set in the same case with a transmitter. The dynamotor and controls are external, connecting to the set through cable bundles. The antenna circuit uses a coaxial plug and transmission line. The receiver cases are relatively poor shields because of the discontinuities existing between case and chassis. Interference gains access to the internal parts of the receiver through these discontinuities, and through the power and audio leads.

Interference input paths and the attenuation of interference through these paths vary considerably with receiver types. These variations are due to:

(a) quality of the case and radio frequency component shielding,
(b) filtering of the input power circuits,
(c) internal routing of interference-carrying wires, and
(d) filtering of radio frequency filament circuits.

Tight case shielding containing as few holes as is practical for cable outlets, and thorough internal shielding of the radio frequency and antenna circuits will lower the receiver interference susceptibility. The "A" receiver is an excellent example of current design incorporating features for interference reduction, as described under Case Shielding, Paragraph 3.4.1.5.

3.4.1.1 ANTENNA AND LEAD-IN

Even though interference from all other possible paths of entry could be eliminated, the antenna would still provide a serious means for coupling interference into aircraft receivers. Poor placement of the antenna or antenna lead-in with respect to the ignition system interference field, the radiation field of a radar set, or of the components of its modulating pulse, may cause large induced voltages which severely affect the output of the receiver. Peak field strengths of airborne radar transmitters, in the vicinity of a communications receiver antenna, may be of the order of hundreds of volts per meter. Under such severe conditions, interference may be introduced into the receiver case over the antenna lead-in or where poor shielding and bonding exist. The antenna and its lead-in provide an entry path capable of causing severe interference in the output despite the frequency discrimination of the receiver. The interference energy may be so intense as to saturate the first stage of the receiver. This type of interference is caused primarily by low-frequency radar transmitters with non-directional characteristics. Radar transmitters

operating at higher frequencies are generally highly directional and seldom cause interference of this type.

The use of a short-shielded antenna lead-in wire is perhaps the best design practice for preventing interference from entering the first tuned circuit of the receiver due to antenna lead-in interference coupling. Amplitude limiting and suppression of the receiver during the radiation period of the radar pulse are other used methods. The method of approach will be determined in accordance with the established purpose of the receiver, although it is best to eliminate interference at the earliest possible point in a receiver. Appropriate location and orientation of the antenna is the only corrective measure available to reduce interference pick-up by the antenna itself.

When the antenna lead-in wire is short, about one foot, and is connected to a shock-mounted antenna coupler, it presents the problem of a semi-rigid connection. The stress due to shock mounting could result in failure of the lead-in wire. However, the lead-in should be shielded as well as short in length and the problem of preventing a break due to stress is left to the installation engineer.

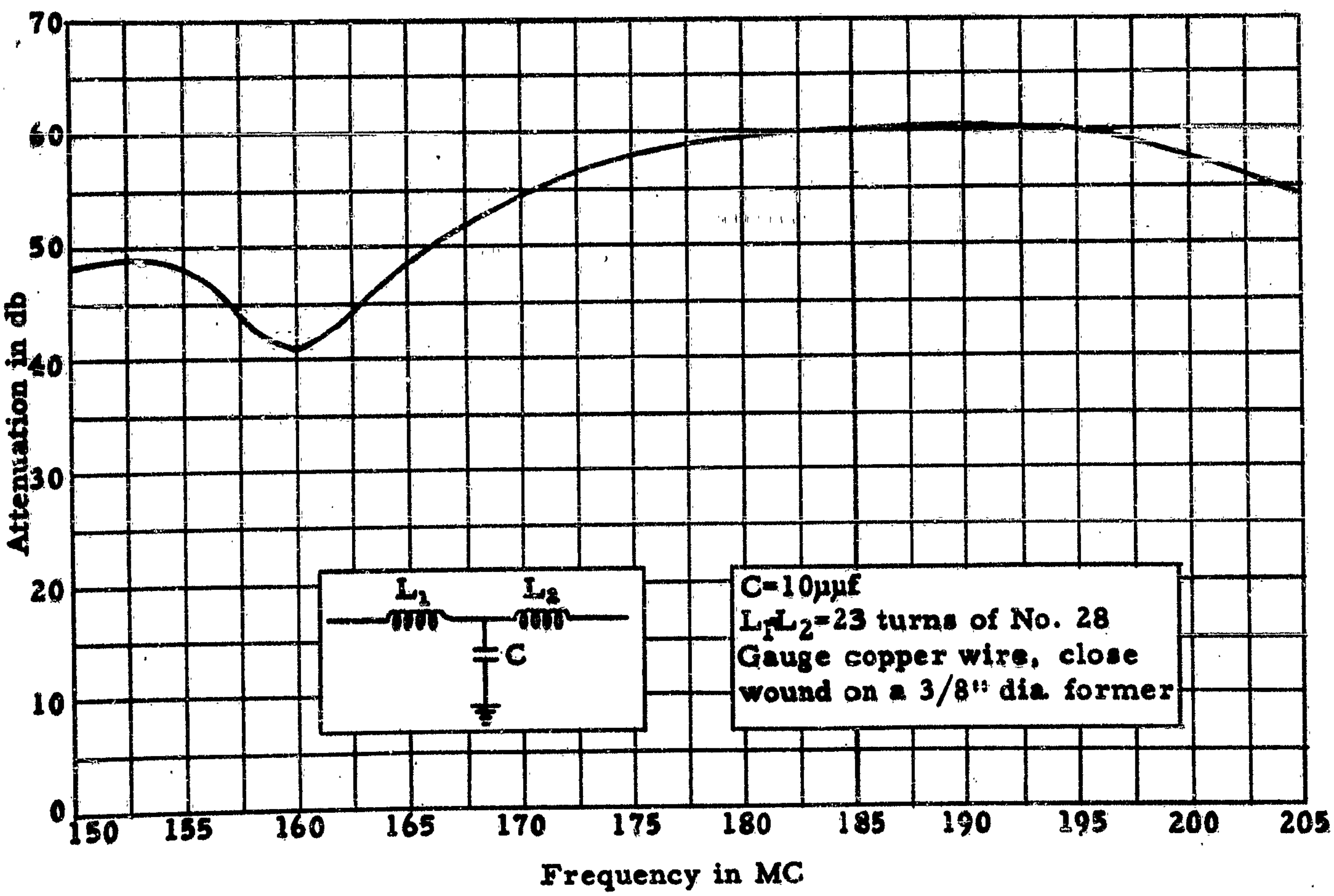


Fig. 3.4.1.1-A Antenna Filter Network for Suppressing Radar Interference in a Representative Communications Receiver

High rejection should be designed into the receiver to provide maximum freedom from interference frequencies other than the reception frequency to which the

receiver is tuned. When high level interference signals are present filter units or wave traps tunable to the radar frequencies may be designed and connected into the antenna lead-in wire to suppress unwanted frequencies and at the same time offer a negligible amount of attenuation to frequencies within the pass-band of the receiver. Filtering is effective in preventing the condition of free oscillations caused by impulse-type interference, for it attenuates the radiated pulses before they can shock-excite the first tuned circuit. An example of this type of filter used in the antenna circuit is illustrated in Figure 3.4.1.1-A. The attenuation frequency bandwidth was sufficient to prevent interference due to sideband frequencies and carrier frequency variations of the radar.

Wave traps when used must be installed at the receiver antenna post or within the receiver case. Such wave traps or filters may consist of parallel resonance choke coils or simply a quarter-wave stub. Choke coils are generally inserted in series with the receiver antenna or grid leads. Quarter-wave stubs are generally connected between the receiver antenna post and receiver case with the far end open-circuited. This provides satisfactory protection against very high or ultra-high frequency interference below 600 mc. The stub is cut to quarter-wave resonance at the frequency to be attenuated. Twisted wire pairs or parallel wires are convenient and practical for constructing the stubs. Wave traps made of a resonant stub using concentric transmission line will have a sharp cut-off characteristic and are not desirable for use against interfering energy coming in at a frequency of a few hundred megacycles.

3.4.1.2 POWER LEADS

All primary power wiring in an electrical system is connected to a common bus bar. Interference sources always impress some portion of their output on the power wiring connected to them, unless completely filtered. Therefore, radio interference originating in the electrical and electronic equipments of the aircraft, except the portion attenuated by the wiring, appears at the input to the receiver. Efficient removal requires knowledge of the frequency of the offending voltages. This information may be obtained by means of a calibrated receiver and probe, noting, however, the limitations pointed out in Appendix V. Protection against conducted interference over the power leads can best be controlled by suppression at the interference source and through good internal receiver design. When these means are improperly employed some external controls are necessary. If the interfering frequencies fall within the band-pass of the receiver, a filter may be constructed for their removal, and inserted in the power line at the receiver plug.

Rotating equipment, radar, IFF and similar gear installed in a modern aircraft produce impulse-type interference which may readily be conducted to the receiver by means of the power line. Network filters used in aircraft at the present time and applied to the power line for attenuation purposes cover a frequency range of approximately 0.2 mc to 30 mc, but the range can easily be extended. The design procedure to be followed in constructing appropriate filters is found under Paragraph 3.1.1.2.

An example of a recently designed filter network capable of attenuating frequencies by at least 40 db in the long and medium wave broadcast and television bands is shown in Figure 3.4.1.2-A. Its declared purpose is to suppress interference

from electrical appliances.

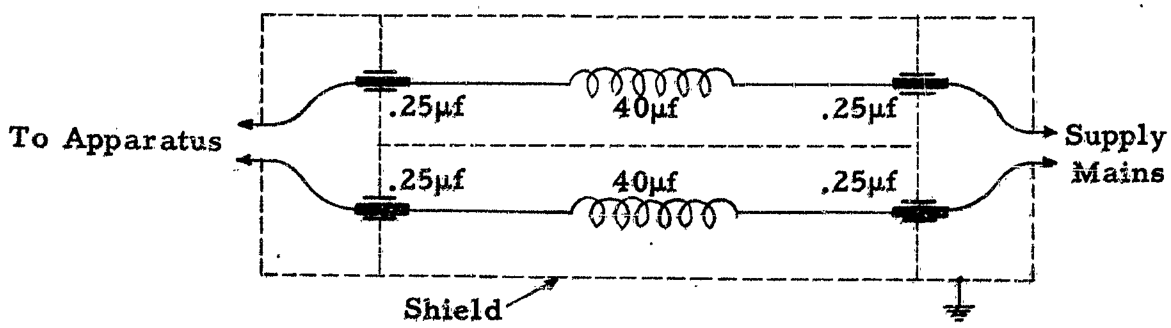


Fig. 3.4.1.2-A Power Line Filter

The filter consists of a single section connected in each lead of the power cable. The series inductor has a value of 40 µh and consists of 60 turns of No. 14 gauge enameled copper wire wound on a cylindrical paxolin form 1-1/2 inches in diameter, with 12 turns per inch. The parallel capacitors each have a value of 0.25 µf and are a special feed-through type. The capacitors function satisfactorily at frequencies considerably above 150 kc since they are designed to have low intrinsic inductance and, therefore, do not become self-resonant until very high frequencies are reached (about 100 mc). At the self-resonant point attenuation is more pronounced since it acts as a series resonant line to ground. Beyond this point the capacitors gradually lose their efficiency as the frequency is increased due to lead inductance. The study and application of feed-through capacitors is outlined in Paragraph 3.1.1.5.

The particular filter capacitor indicated here has a test voltage of 1500 volts, and is acceptable for use on 220 volt AC or DC supplies. Similar capacitor types are also available with a test voltage of 2250 volts. The manufacturers rating on the central conductor is fifty amperes up to a temperature of 85° C. It functions here as a fifteen ampere filter, and under full load the measured temperature increase did not exceed 40° C.

When filtering is done at the receiver the filter sections are reduced because the range of frequencies to be attenuated is only that to which the particular receiver is susceptible. There is some attenuation of interfering currents while traveling from source to receiver, especially when the distance is appreciable. However, a recent study of noise propagation and distribution has shown that very little attenuation exists in a wiring system.

A great deal of experimentation concerning the use of a lossy line, or lossy dielectric materials and their application to power cabling is now going on. Experimental data gathered so far indicates that lossy, low dielectric constant materials incorporated in practical filter design is capable of greatly reducing interference. Higher capacities and losses may be obtained by using ceramic dielectrics. However, the effective application of higher dielectric ceramic materials is presently limited because of the difficulty of shaping the forms (limited to fairly thick-walled tubing and flat sheets). Several low dielectric materials have been tested, of which shellac appeared to be the best. Filters using lossy materials are discussed in

Paragraph 3.2.1.

3.4.1.3 CONTROL CABLES

Electrical or mechanical control cables attached to a receiver, even though not connected to a source of interference, can readily act as an antenna, picking up interference by induction or radiation and transfering it to the receiver. Once they pass beyond the screening effect of the receiver case, there is nothing to prevent the interference from affecting the receiver output. There are many cables and lead wires required in a modern aircraft, frequently routed along a common path. Those carrying pulse currents have little difficulty in injecting a portion of their energy into receiver lead wires, and consequently causing interference. It is worth recalling that an overhead waterpipe or steel beam with radio-frequency energy induced in it from a nearby transmitter can change from a normally good ground to act like a long wire antenna. For example, at 28 mc there will be a high voltage point approximately every eight feet; that is, it will have standing waves on it and it will radiate. Any control cable greater than one-eighth wave length may, in a similar fashion, act as an antenna for its resonant frequency, and hence, there is some possible interference frequency at which it acts as an antenna.

This interference is eliminated by effective shielding of the control cable. Actually all leads carrying pulse currents should be shielded and routed separately. However, no interference path may be neglected. The control leads should be grouped in a separate amphenol connector which incorporates an internal grounding ring for better grounding of the shields to the chassis. Feed-through capacitors located in the connector will help suppress very high frequency interference.

Electrical control cables which terminate in the receiver and are run in a group of wires, may introduce interference signals into the receiver wiring through inductive or capacitive coupling. When feasible, isolation of cabling is effective; however in some cases filters are required. The filter should be applied at the point of entry of the control cable to the receiver and may easily be incorporated in the original receiver design. Usually it is a low-pass filter, and the frequencies to be filtered out may be ultra-high, very-high, or medium-high frequencies depending on the frequency range of the receiver with which it is to be used. For this reason it is desirable to include it in the original receiver design. The low-pass filter will not attenuate interference conducted to the receiver on the fundamental frequency of a radar transmitter. This will require a line filter of ultra-high frequency design installed within the receiver. These may be series-resonant mica capacitors or sparkplates of the type discussed under ignition interference in Paragraph 3.2.3.1.

3.4.1.4 AUDIO OUTPUT LEADS

Though designed to carry audio intelligence, phone leads may also act as a pick-up for interference. Power line frequencies that gain access to the audio output circuit over the phone leads may be amplified sufficiently to become audible in the headset. The receiver output circuit should incorporate design features that will minimize reverse coupling through the receiver stages and prevent amplification of unwanted audio frequencies. Phone leads must be kept separate from any wiring carrying alternating current, and the compact construction of aircraft receivers may make it advisable to shield the audio system from the radio-frequency section within

the receiver case.

Once interference signals enter a receiver, they gain access to amplification stages in various ways and with varying degrees of attenuation. Tuned circuits may be susceptible to coupling with filament or relay wiring. Inadequate case and inter-stage shielding are coupling paths which also increase susceptibility. Production methods of cabling provide tight coupling between interference-carrying wires and high impedance input and sensitivity control circuits. Receivers employing band switching appear to be more susceptible because of inadequate shielding between radio frequency coils.

The radio interference signals that gain access to the receiver case must eventually reach the last RF stage or second detector in order to adversely affect the receiver output. This internal interference path can generally be eliminated by proper design of the output stage. Reverse coupling either through the tube itself or through the associated circuitry, can be greatly reduced by careful selection of circuit components and internal stage shielding.

If the interference injected into the audio leads is at radio frequencies, a simple by-pass capacitor is usually sufficient to prevent its entry into the receiver. Satisfactory operation is usually achieved by installing a 0.01 μf capacitor, or smaller value, between the audio "hot" lead and the receiver case at the point where the lead enters the receiver. This capacitor offers a high impedance to audio frequencies, but presents a low impedance path to the radio frequency energy.

The use of filtering as well as shielding of the phone leads is sometimes necessary to prevent radio frequency interference and any extraneous audio disturbances. Shielding is accomplished in the conventional manner; a metallic sheath covering the phone leads, and the sheath well grounded to the receiver case.

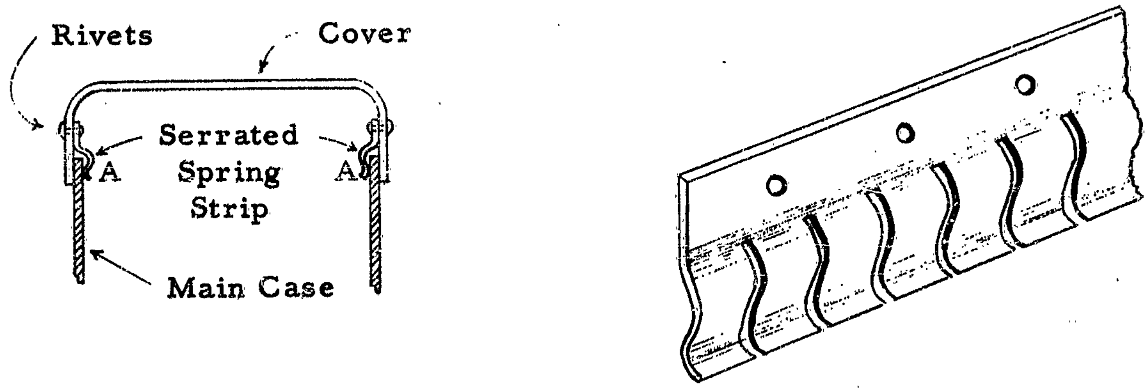
3.4.1.5 CASE SHIELDING

The primary purpose of the receiver case must be to shield the receiver from any external interference fields. The number of mechanical discontinuities must be kept to an absolute minimum, and those that are required must be electrically continuous across the interface. A multiple point, spring loaded contact is a very efficient method of obtaining electrical continuity.

Bonded screening of suitable conducting material must be used to cover all louvres and other apertures used for ventilation. No. 8 mesh should be sufficient for screening out most frequencies used in present-day equipments.

The multiple point, spring loaded contacts mentioned could be constructed in a number of ways. In general, a serrated shim inserted in the aperture of the discontinuity will be satisfactory. The serration gives enough spring pressure at its points of contact for electrical continuity, and no spring pressure is required at any other point on the shim. A sketch of a spring joint and a serrated spring is shown in Figure 3.4.1.5-A. The materials used in constructing the shims could be beryllium copper, german silver, phosphor-bronze, sheet steel, or tempered aluminum. The receiver case, however, should be constructed of the same material to prevent corrosion and resultant electrical discontinuity. If the materials are different, the

shim must be protected by cadmium plating or alternative methods providing good electrical contact while preventing corrosion. This practice likewise conforms with service specifications.



(A) Serrated Spring Joint

(B) Serrated Spring

Fig. 3.4.1.5 Multiple Point Spring Loaded Contact Joint

Receiver plugs must be free of paint or varnish between plug shell and receiver case, as well as between the plug and its shell. Such non-conducting materials at these points produce electrical discontinuities which could be a source of serious interference to normal receiver operation. The shield ground likewise must extend completely around the plug.

Mention has been made of tests conducted on representative communication receivers to determine their susceptibility to interference. Of these Receiver "A" is the best approach to a low-interference installation. The receivers are compact, ~~individual units~~ carrying the dynamotor outside the shielding, and are housed in aluminum covers. The covers are fastened down to the chassis and front panel almost continuously with screws. The radio-frequency sections are very well shielded on the interior. Serviceability is sacrificed in favor of interference reduction. However, a large percentage of the present airborne communication receivers have poor case shielding which will not effectively attenuate radar energy when it impinges on the receiver case. The case shielding could be decidedly improved in design by providing that all joints in the case make multiple contact around their peripheries. The serrated spring shims, as previously mentioned, could be installed between joints, and all louvres or openings in the case covered with copper screening effectively bonded to the case.

Shielding effectiveness also hinges on the thickness of the shield and the type of material used. The depth of penetration of interference currents in the metal wall is discussed in Paragraph 3.1.2. Considerable attention is being given the development of effective shields at the present time. See Appendix XI. An excellent degree of attenuation is a definite possibility using thin-walled ferrous metal shields of high effective permeability, and plated with non-ferrous metals of high conductivity. Plating the shield with a non-ferrous metal increases its reflection loss.

3.4.2 INTERNAL COUPLING

The most direct coupling path, and one having least attenuation, is through the radio frequency filament circuits. Interference entering receivers through the power input leads, branches directly into the filament wiring. Several tests made on representative receivers, showed an average attenuation of 5 to 10 db measured between the input plug and first radio frequency filament. Further, the first radio frequency stage filament was isolated from the power supply and separately heated. When the signal was fed into the isolated filament, the receiver output was essentially the same as with normal operation. However, keeping the first radio frequency filament isolated, and applying the signal to the remaining filaments, the outputs of Receivers "B" and "C" were down 10 db while the Receiver "A" output was down about 30 db. Tests were also made with first and second stages inoperative. Tabulated results showed that approximately 90 percent of the interference response in a receiver is due to excitation in the first radio frequency stage, and that interference entering a receiver through the power input leads is largely coupled into the first radio frequency stage through the filament-grid capacitance. This capacitance was found to be approximately 5 to 6 μμf. Figure 3.4.2 shows the results of the described tests for typical HF command receiving equipment (Receiver "A").

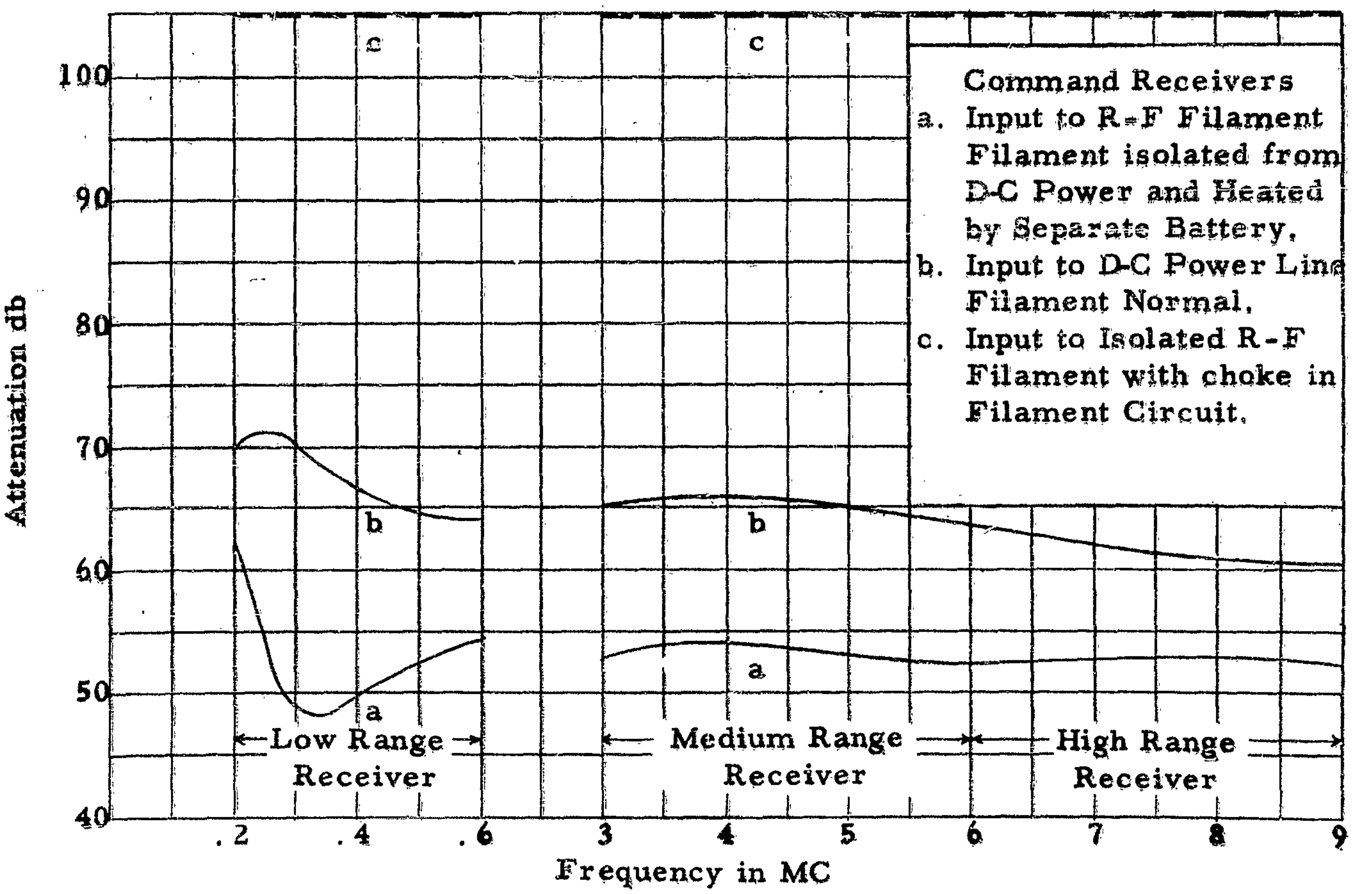


Fig. 3.4.2 Attenuation with Reference to Antenna Input Voltage

3.4.3 SPECIAL CIRCUITS

Where special circuits are devised for coping with a certain type of interference,

they may be incorporated in any receiver if that receiver is susceptible to that interference type. For example, a gated amplifier is used to operate a radar receiver only during selected intervals of time. This same circuit may be used to block a counting circuit during an interfering pulse, as in the case of an altimeter. Such pulses may also be used as blanking pulses to turn off the beam of a cathode-ray tube, in order to eliminate a horizontal retrace line. A basic circuit performs a basic function. The circuit may be used wherever there is need for that function.

3.4.3.1 RADIO RECEIVERS

The tuned circuit of a radio-frequency input stage is essentially a band-pass filter. It usually has the characteristic of high selectivity, passing the frequencies within its acceptance band, while rejecting all others. When unwanted signals or interference, comparable in amplitude to the desired signals, appear in the frequency ranges adjacent to the pass band, they are effectively attenuated by the tuned circuit. Succeeding stages will further attenuate the minor amount of interference which does pass through the first stage. However, if the interference is very strong, of the order of volts as it may well be from a neighboring highpowered radar unit, the received impulse will produce ringing of the tuned circuit at its natural resonant frequency. This effect is passed along and amplified in conjunction with the normal signals, and the interference will appear in the output. The same is true of extremely strong radiation from a radio transmitter operating on an adjacent channel. Even though the transmitted energy is attenuated by the tuned circuit of the receiver, the response is still high within the pass-band, and the familiar effect known as cross-talk occurs in the output. Therefore, it is necessary to provide for interference reduction by design techniques.

By using more than one radio frequency stage, additional selectivity may be provided between the antenna input and the converter unit. It is possible to increase both selectivity and fidelity by adding more stages, in cascade. This involves broader tuning of the individual stages to avoid a loss of fidelity by using compact coils wound with relatively small wire. The resultant reduced Q causes a loss in gain which is offset by the additional number of stages. A simple arrangement for increasing selectivity, by means of an extra tuned circuit, is illustrated in Figure 3.4.3.1-A.

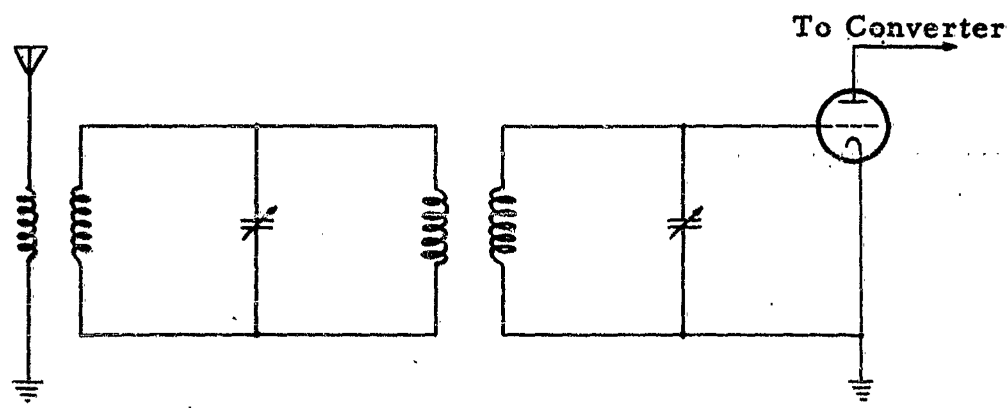


Fig. 3.4.3.1-A Additional Tuned Circuit Preceding Converter

Such circuits will also provide isolation of the oscillator stage and help prevent

radio frequency energy at the oscillator frequency from reaching the antenna and causing undesirable radiation. However, they do not satisfactorily suppress interference consisting of strong pulses. A wave trap in the antenna circuit is required to properly eliminate strong interfering pulses. This is pointed out in Paragraph 3.1.4.2.

Limiter circuits may be incorporated in receivers with the following limitations:

(a) They are suitable only for use in receivers which have some radio frequency gain and rectify at a fairly high level, at least 0.1 volt.

(b) Limiters are useful for suppressing low-pulse recurrence frequencies below 100 times a second, but are less effective for suppressing those up to 500 times a second.

The length of an interfering pulse may be considerably increased as it passes through successive stages of a receiver. This is a direct result of the tuned circuits "ringing" when subjected to pulse excitation. (See Paragraph 1.7.2.) The time, t_o, in microseconds which it takes for the amplitude of the pulse to fall to approximately 4 percent of its initial value is given by the reciprocal of the bandwidth in megacycles of the tuned circuit:

$$t_o \text{ (microseconds)} = 1/(\text{Bandwidth mc})$$

The ringing waveform in combination with the local oscillator frequency will then produce a ring at the intermediate frequency of the receiver and appear in the output.

A circuit required to pass short pulses without distortion must be able to pass a wide range of frequencies. For example, a video pulse amplifier must have a reasonably flat frequency response up to high frequencies. If the bandwidth of a tuned circuit is large, the pulse will not be lengthened to any marked extent, and, because the pulse is of short duration, the limiting action will be much more effective. The pulse that is superimposed on the desired signal, is prevented from reaching the audio section by the action of the limiter. At the same time there is negligible distortion of the desired signal due to the short time interval of the pulse as compared to the period of the signal.

However, unavoidable interference power in the output of an amplifier increases in proportion to the bandwidth. Furthermore, the gain per stage in an amplifier is, in general, inversely related to bandwidth, so that for a given overall amplification a broadband amplifier requires more stages than one with a narrower band. It thus becomes important to judge the best bandwidth for a particular application.

Pulses with low repetition rates are adequately suppressed by limiters while higher recurrence rates would cause a proportionately larger amount of interference at the output terminals. Thus low repetition rate and a large bandwidth represent ideal conditions under which to operate limiters.

The second detector is the first part of a receiver where an amplitude limiter can effectively be placed. The radio frequency voltages at the input to the receiver

are too small to operate any known forms of limiter or rectifier. Usually the interfering pulse at the second detector has not become too long for effective limiting. Pulse lengthening is a function of initial energy and the bandwidth of intervening circuits. Limiters will require less operating time with low initial amplitudes of an interfering pulse and greater bandwidth in the tuned circuits. The duration of the pulse at the point of limiting is a very important factor because a portion of the desired signal is affected every time the limiter comes into action.

Limiters are primarily restrictive devices and distortion will result from their use, particularly when the input exceeds the limiting threshold. Limiters of the instantaneous interference-peak type generally distort the output whenever the modulation of the incoming signal exceeds a definite percentage. The distortion effects can be intensified by the transient distortion characteristics of the audio amplifier. In general, it is desirable to use triode tubes in the audio amplifier or degenerative feedback sufficient to prevent oscillations because of insufficient damping of the output circuit.

When both a modulated carrier and pulses of interference are present at the output of the final intermediate amplifier, then the current through the audio frequency output resistor is of the form shown in Figure 3.4.3.1-B.

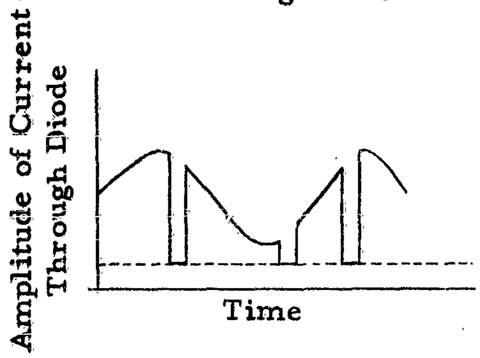


Fig. 3.4.3.1-B Current Through Limiting Diode Showing Limiting Action on Interfering Pulses

If the cut-off point is adjusted to coincide with the negative peaks of the audio waveform then, as observed in the figure, maximum attenuation of the interfering pulse would occur with minimum distortion of the audio waveform. Under the ideal condition of wide bandwidth, as is commonly encountered in very high frequency receivers, a series limiter provides an average attenuation of 30 to 35 db of the unwanted pulse. These limiters also offer some degree of protection in receivers operating in the lower frequencies and at their widest possible band acceptance. Amplitude limiting gives a valuable degree of protection against atmospheric and ignition interference.

Parallel limiters are not as effective as series limiters, and under the same ideal conditions mentioned previously, provide an attenuation generally of about 20 db of the interfering pulse. Furthermore, they do produce distortion of the audio waveform since they reduce the output by shunting action, but are not a complete short circuit. The time, t_o, during which distortion occurs, varies inversely with the percent of modulation. In general, it is less than 0.1 millisecond above 50 percent modulation, as may be observed in the typical graph of Figure 3.4.3.1-C.

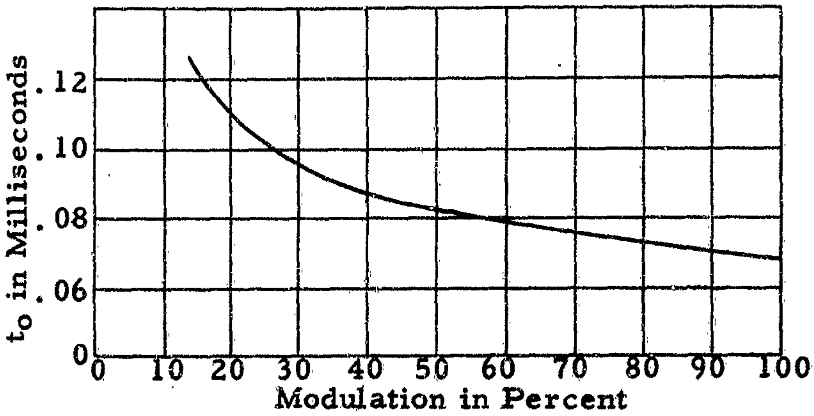


Fig. 3.4.3.1-C Distortion Interval of Audio Waveform Measured Against Percent Modulation

There are cases where interference pulses with frequencies of the order of 200 megacycles penetrate the receiver case or enter through external leads and are internally coupled to the audio frequency amplifiers at sufficient amplitude to result in grid-circuit detection of the pulse. In such cases satisfactory suppression is obtained by the use of a resistance-capacitance decoupling network. This combination has very little effect upon the audio frequency signals, but will greatly attenuate the interfering energy due to the low input impedance of the grid circuit at high frequencies. A typical circuit is shown in Figure 3.4.3.1-D.

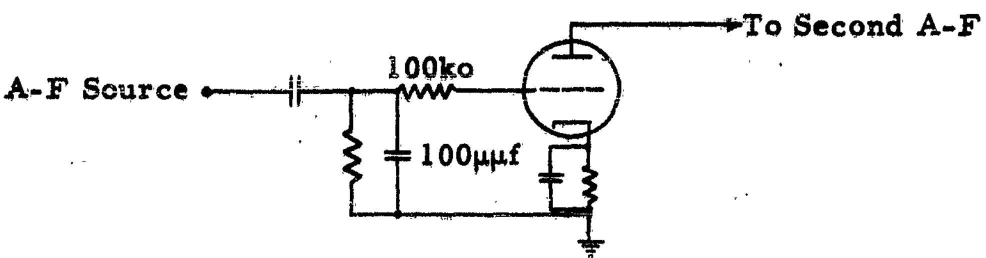


Fig. 3.4.3.1-D R-C Decoupling Network at Input to First AF Stage

It is quite possible for the interference entering through the amplifier case or external leads to the amplifier to be present in any part of the amplifier circuit. This necessitates precautions in each stage of the amplifier. Generally the first stage is most important since greater gain results in subsequent stages. The blocking resistor of the resistance-capacitance network is in series with the grid and located as close as possible to the tube terminal. The use of a short connecting lead wire will minimize interference injection between the series resistor and the tube. This combination is similar to an L-type filter with the inductance replaced by a resistance. The resistance should be designed to be much greater than the reactance of the capacitor for the interference frequencies. A ratio of 10:1 is useful in practice. This permits most of the interfering energy to be shunted to ground.

Resistance-capacitance networks are usefully employed as decoupling networks in the plate circuits of an amplifier to prevent interstage coupling and possible oscillations. The voltage output obtained from a common B supply is not fixed but varies with the current demand. Also, any ripple appearing at the output of the

power supply filter is impressed on all the amplifier grids, except the first. When using a decoupler the voltage across the condenser is very nearly constant and independent of any power supply variations. The networks act independently of one another and thus isolate the stages. A cascade arrangement, where the first stage possesses the largest amount of decoupling, is illustrated in Figure 3.4.3.1-E.

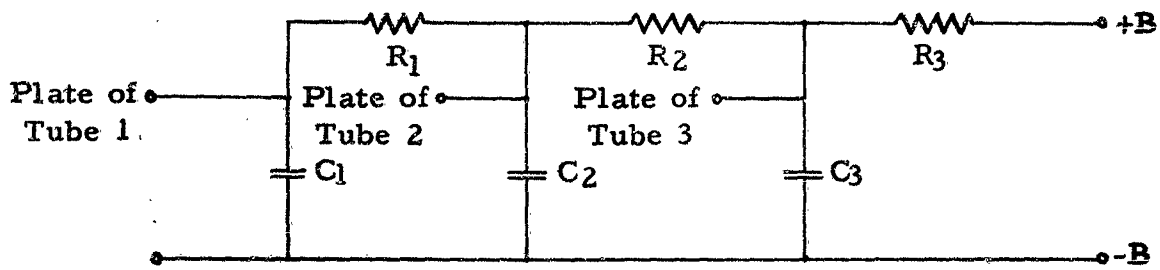


Fig. 3.4.3.1-E Decoupling Networks in Cascade

3.4.3.2 RADAR RECEIVERS

The problem of interference rejection in radar receivers is somewhat different from that encountered in communication receivers, since radar interference is nearly always the result of signals occurring so close to the radar frequency that very little can be done by improving receiver selectivity. The ability to deliver intelligible information to the radar indicators in the presence of on-frequency interference gives a better indication of the radar's quality than its selectivity in the usual sense of the word, i.e., the ability to reject disturbances at frequencies other than the desired signal frequency.

Most receivers have sufficient amplification to give appreciable interference outputs in the absence of a received signal, for which reason the sensitivity of a radar receiver is nearly always determined by its susceptibility to interference. Furthermore, amplitude-modulated interference is of greater concern in radar receivers than frequency-modulated interference because amplitude modulation envelopes approximate rectangular pulses in shape. Since adequate detection of rectangular envelopes depends more on receiver bandwidth and phase shift than on linearity of amplitude response, non-linear amplitude response in the form of limiting is permissible in most radar receivers. The limiter suppresses the amplitude-modulated interference and produces a visual output relatively free from interfering signals, even though the frequency-modulated interference actually may be increased by the action of the limiter.

Three main types of receivers are used in radar applications which differ in regard to the frequency region within which most of the necessary signal amplification results: (1) superheterodyne receivers which convert the modulated radio frequency signals to an intermediate level, at about 30 megacycles, before amplification; (2) super-regenerative receivers, which use a regenerative radio-frequency amplifier, with the oscillations quenched in a time interval about equal to a pulse width; and (3) crystal video receivers, which detect the modulation signals and amplify the resulting video signals. The majority of microwave radar receivers are of the superheterodyne type since this permits the largest amount of amplification to

take place in a fixed-tuned amplifier.

At frequencies below the microwave region, the first detector may be preceded by one or more stages of radio-frequency amplification. The crystal mixer may be replaced by a vacuum tube mixer. The additional gain obtained from the radio frequency stages results in better image rejection, improved signal-to-interference ratio, and reduction of the radiation of local oscillator power.

Most airborne radar systems today operate in the microwave region, and mixers for use at frequencies higher than 3000 mc are usually of the waveguide rather than coaxial type. At these frequencies the local oscillator contributes rather serious interference energy which can be sharply reduced by using a balanced mixer. One form of balanced mixer is the waveguide magic-tee, as illustrated in Figure 3.4.3.2-A, with the crystals installed in the arms of the waveguide, parallel to the electric field.

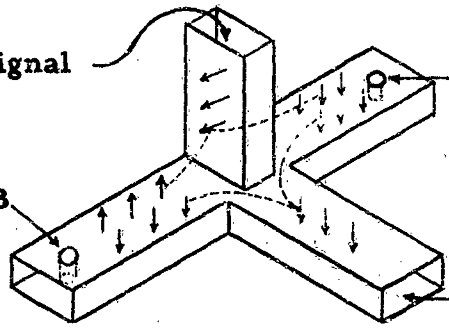


Fig. 3.4.3.2-A Magic-Tee Balanced Mixer

The arrows in the figure indicate the direction of the electric field in each arm. A waveguide type of mixer divides the local oscillator power equally between the two arms containing crystals A and B and prevents energy transmission out of the signal input arm. A simple and rather obvious explanation for such a characteristic follows directly from the geometrical arrangement of the four arms. The arms containing the crystal mixers and the local oscillator input arm form a shunt tee and provide for the continuity of the local oscillator electric field between the arms. Similarly, the crystal arms and the signal input arm form a series tee and provide for the continuity of the signal input magnetic field between the arms. Thus the local oscillator and the signal input powers divide equally between the crystal arms when the impedances of these arms are matched. However, since there is no provision for continuity of either electric or magnetic components of the transverse electric wave between the signal input arm and the local oscillator arm, no energy can be transferred from one to the other. Therefore, local oscillator radiation is minimized through the use of a magic-tee balanced mixer.

A crystal mixer has a gain of less than unity, making it mandatory to control the interference developed at the input stages of the intermediate-frequency amplifier. Triodes are preferred to pentodes in these stages because of their lower interference characteristics, but they cannot always be used because of their large input capacitance which leads to extremely poor performance at frequencies as high as 30 megacycles. In the circuit shown in Figure 3.4.3.2-B triodes are used in the intermediate

frequency 30 megacycle amplifiers. The triode-connected 6AK5 vacuum tube has a load impedance of about 200 ohms presented by the cathode of the second tube. This low impedance secures stability of the first tube, whereas in the second tube stability results from the grounded grid, shielding the input (on the cathode) from the output. The neutralizing coil between plate and grid of the first tube confers extra stability. It helps to prevent the output impedance of the first tube from falling off, and results in minimizing the interference from the second tube. This circuit yields an intermediate frequency interference figure about 2 db lower than a similar intermediate-frequency input employing a pentode.

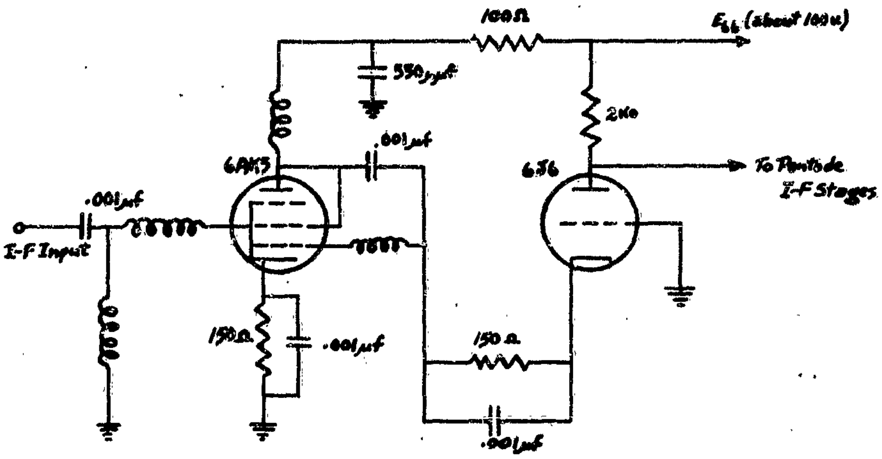


Fig. 3.4.3.2-B Diagram of Grounded-Cathode, Grounded-Grid Dual Triode Input for 30 MC IF Amplifier

Since interference in the output of an amplifier increases in direct proportion to the bandwidth, it is important to restrict the intermediate frequency bandwidth as much as possible. It must be sufficient, however, for adequate transmission of the signal. If a radar system is employed for search purposes, then the visibility of an echo in the presence of interference is the primary measure of performance. In that case the best circuit bandwidth has been found to be 1/T cps, where T is the pulse width. This results in an intermediate frequency amplifier bandwidth of 2/T cps in order to transmit the double-sideband signal at this point. Thus certain fire-control radar equipments, making use of the leading or lagging edge of a received pulse for precise range measurements, require a broader bandwidth to assure a minimum rise time of the displayed pulse and consequently a more precise determination of the position of the pulse. For example, a radar pulse width of about 5 microseconds would require a bandwidth of $1/5 \times 10^6$ cps or 0.2 mc. A short pulse of 1/5 microsecond duration, however, requires a bandwidth of 5 mc.

Consideration must be given to the behavior of the radar receiver in the presence of excessive signal strength, as from neighboring radar or "jamming" signals. Intentional jamming may be of the "window" type, that is, it may be caused by strips of metallic foil which cause numerous fluctuating echoes and thus obscure the presence of aircraft or it may be caused by radio waves, either modulated or unmodulated from jamming transmitters. The most common form of modulation is that by long pulses, termed "railings" because of their appearance on a scope. Accidental jamming may be caused by strong echoes from land targets, rough water

surfaces, or clouds. It may also be caused by other high frequency equipments, and if so will appear either as continuous wave or as "railings" jamming.

The main purpose of anti-jamming circuits is to prevent receiver saturation. Manual adjustments are impractical because it is quite impossible to follow the rapid fluctuations in jamming with a manual control. A signal of 80 db above normal receiver interference level coupled with 50 db of clutter would appear as a 30 db signal if the receiver gain were properly reduced. With normal gain, however, the signal would be invisible since the clutter would saturate the receiving system. This is particularly true in the case of an intensity-modulated indicator, such as a plan position indicator. When the beam intensity is increased too much the focus is destroyed, and the spot is said to "bloom". The plan position indicators have a small dynamic range, around 10 to 20 db, and therefore require a limiter stage in the preceding video amplifier. This prevents strong signals from causing blooming.

Four types of circuits are useful against jamming and clutter:

(a) Sensitivity time control circuit. These circuits control the receiver gain as a function of time after the initial radar pulse. The gain is reduced when the radar pulse is first sent out and then gradually increased to normal value as determined by the time constants of the receiver. Since gain is made a function of distance, this circuit is useful only when a desired echo is greater in amplitude than the interference echoes at all ranges, and when their ratio is maintained for increasing ranges. This is possible only when the interference source is a target without strong directional characteristics, such as sea or land surfaces.

(b) Automatic gain control circuits. An instantaneous automatic gain control rapidly decreases the gain of an intermediate-frequency stage when the stage output increases beyond a value determined by the circuit constants. This action prevents stage saturation. It is advisable to protect the last two or three IF stages with this type of control. The circuit usually operates with a time constant of about 20 microseconds.

(c) Short-time-constant networks. The coupling between the second detector and first video stage is provided with a very short-time-constant network. The network serves to remove or attenuate, by differentiator action, the DC and low-frequency components encountered in continuous wave or low-frequency modulated jamming. The time constant is usually made equal to the radar pulse width.

(d) Bias-control circuits. These circuits automatically supply a bias to the second detector which prevents the high-frequency components of interference-modulated jamming or clutter from saturating the video section. The circuit can be designed with a short-time constant. For this reason a delay network is also necessary so that individual signals will not be reduced in amplitude, i.e., cut off too soon.

The short-time-constant and the bias-control circuits are most effective when used in conjunction with the instantaneous automatic gain control circuit. At frequencies below the points where the short-time-constant circuit cuts off, the fast

time constant and instantaneous gain control are an effective combination against jamming by modulated or unmodulated continuous waves. Modulated jamming and most types of clutter, expecially that caused by clouds, are best controlled by a combination of bias and instantaneous automatic gain control.

High power pulsed radar systems operating near each other can readily cause mutual interference, even though there is considerable separation of their operating frequencies. The radar receiver does not offer sufficient attenuation for extremely strong off-frequency signals. Interference may also be caused by pick-up of the large video signals radiated from a nearby modulator. Blanking circuits have offered the most effective solution to this type of interference. A receiver gating pulse is developed and applied to one or more intermediate-frequency grids and synchronized with the transmitted pulse of the interfering set. (See Special Circuits, Gating Circuit, Paragraph 3.1.4.1.)

Receiver gating can be accomplished in the intermediate-frequency or video section. The stage may be cut off by reducing the plate or screen voltages, by making the suppressor or control grid voltages negative, or the cathode voltage positive. When applied to an intermediate frequency stage, the gating pulse does not produce any disturbance at the receiver output, since the amplifier cannot amplify the frequencies contained in the pulse. However, if a video amplifier is gated, the gating pulse produces a pedestal in the output on which the normal signals ride. This occurs because the video amplifier has an appreciable response in the range of frequencies contained in the pulse. When the pedestal is present in the output, it may be removed if necessary. For example, a pedestal is not permissible when signals are applied to an intensity-modulated cathode ray tube since such a tube is not normally biased beyond visual cut-off.

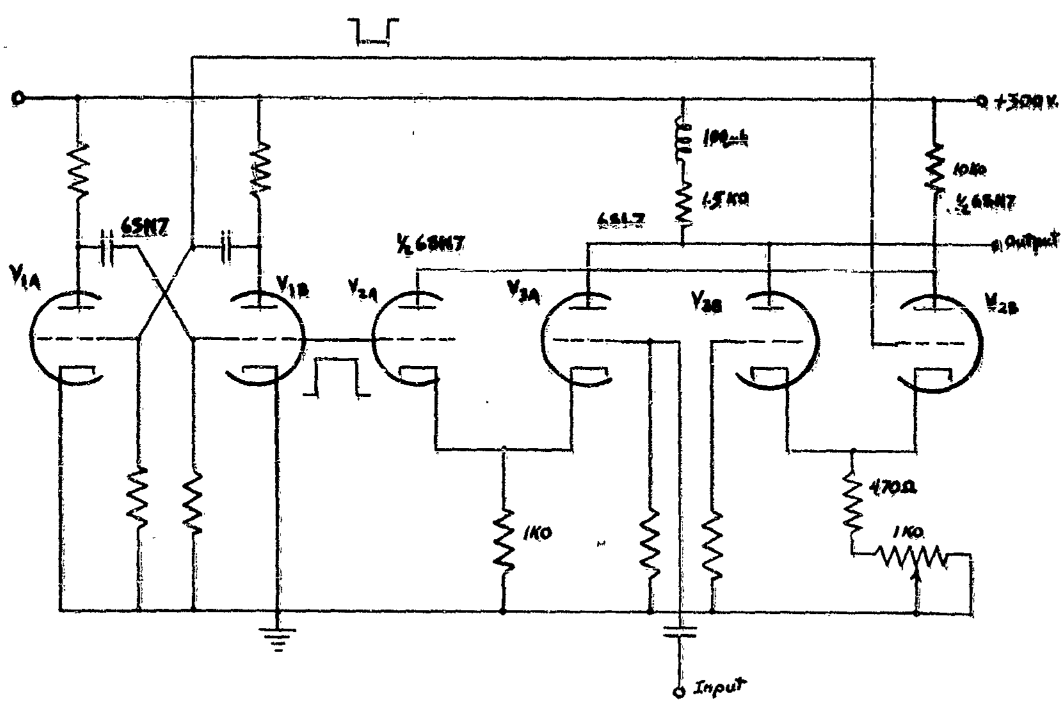


Fig. 3.4.3.2-C Cathode-Gated 6SL7 Video Stage, with no Pedestal in Output

A non-pedestaling cathode-gated video stage is illustrated in Figure 3.4.3.2-C. There is a gain of about unity and sufficient bandwidth to handle pulses of about one microsecond. The stage is capable of handling positive or negative signals of a few volts amplitude. A 6SN7 multivibrator (Eccles-Jordan circuit) supplies the gating pulse. V_{2A} and V_{2B} are cathode followers for driving the 6SL7 cathode, while V_{3B} carries the output plate current when V_{3A} is gated off. When V_{1A} is conducting, the positive potential on the cathode of V_{3B} rises sufficiently to cut the tube off. When V_{1B} is conducting, then V_{3A} is similarly cut off.

3.4.4 DESIGN CONSIDERATIONS FOR MINIMUM OSCILLATOR RADIATION

Receivers must not be allowed to radiate signals whose frequency will adversely affect the basic function of other electronic equipments in an aircraft, or in other aircraft, such as may occur in flight formations. There are several sources of incidental radiation in receivers, but by far the most serious is the high frequency local oscillator radiation of a receiving system. This is particularly true when the receivers operate in the very high frequency range or higher, since the local oscillator frequency is removed from the carrier by only a small percentage. Radiation interference from oscillators can be broken down to two types: antenna radiation and chassis radiation. Existing literature deals predominantly with antenna radiation with little mention made of the problems of chassis radiation. In the lower frequency ranges, where the chassis is small compared to a half wavelength, chassis radiation is no problem. However, when the chassis size is such as to approach half wave resonance at a particular operating frequency, it becomes an efficient radiator. This is true in the case of television receivers operating in the frequency range of 174 to 216 mc. The chassis usually is big enough to approach half wave resonance.

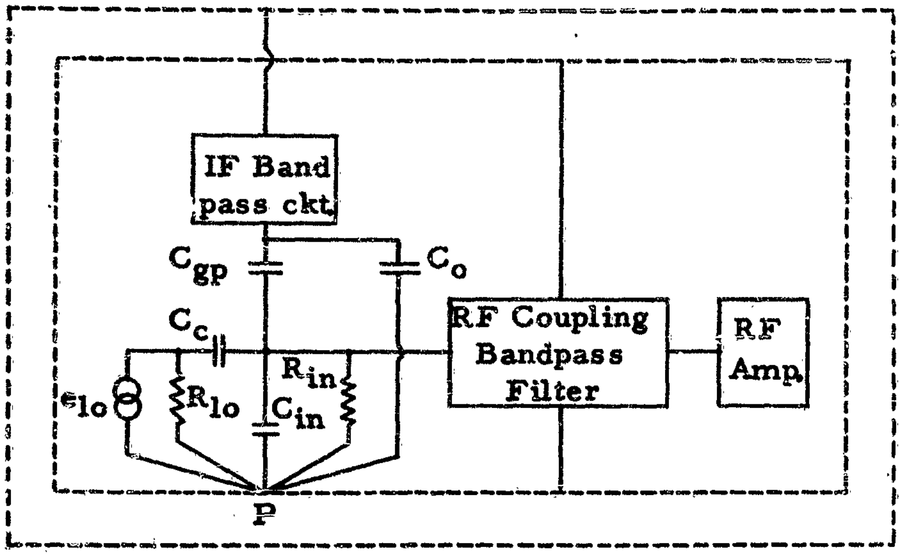
3.4.4.1 CHASSIS RADIATION

Certain measures designed to minimize oscillator radiation are effective for both the antenna and chassis types of radiation. The appropriate technique of shielding will aid in confining the local oscillator energy so that reverse transmission by means of the antenna or lead-in will not occur. It will also prevent the excitation of the chassis, and thus prevent chassis radiation. Triggering of Airways Marker Beacon receivers by television local oscillator radiation has occurred and indicates that this form of undesirable signal radiation does take place. The local oscillator of a television receiver is about 21 mc above the station being tuned in and when channel 2 is used, 54 - 60 mc, the local oscillator frequency is within the region of 75 mc, which is the frequency of Airways Marker Beacons.

The problem of chassis radiation dictates the need to prevent the chassis from becoming excited by the local oscillator energy or to totally reflect the energy radiated if the chassis is excited. A metallic receiver case will aid in reflecting energy which is radiated by the chassis, but the case should be looked upon only as a secondary or outer shield, and emphasis placed on confining this energy to the region of the local oscillator itself. The metal case functions primarily to permit an interference-free region within.

A high degree of shielding is required in the radio frequency section of aircraft receivers for the purpose of reducing their susceptibility to interference. It only remains to extend the principle of shielding to include the problem of oscillator

radiation.



e_{lo} = Open circuit local Oscil. Voltage
R_{lo} = External damping on local oscil.
C_c = Coupling reactance to mixer grid
C_{in} = Input cap. of mixer
R_{in} = Input resistance of mixer circuit and tube
C_{gp} = Mixer grid plate capacitance
C_o = Mixer output capacitance

Fig. 3.4.4.1-A Shielded Local Oscillator Equivalent Circuit

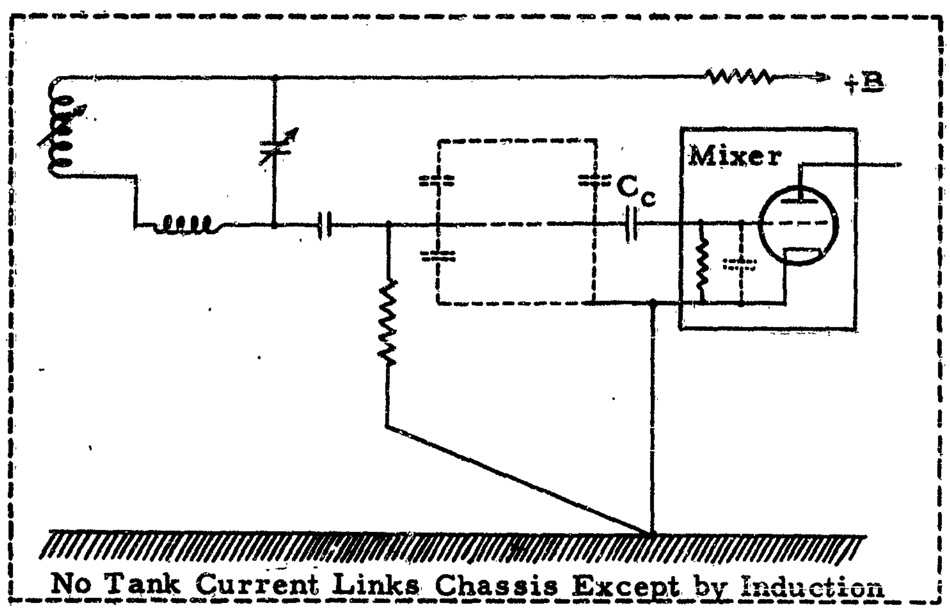
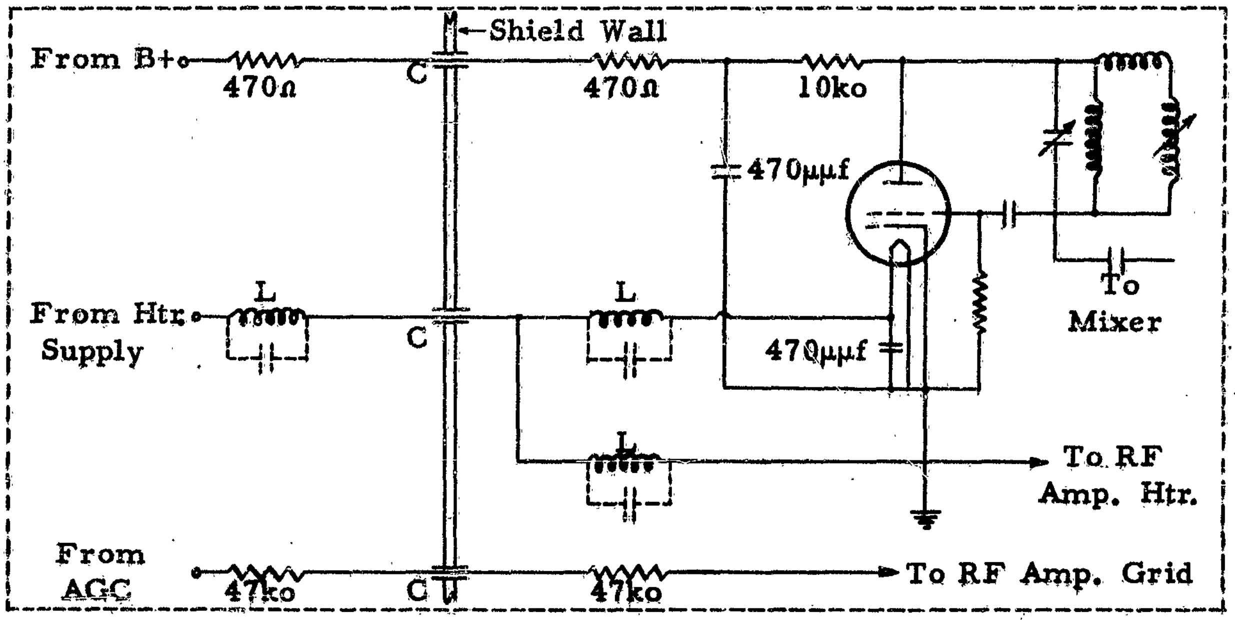


Fig. 3.4.4.1-B Good Oscillator Design Reduces Chassis Currents

Figure 3.4.4.1-A illustrates an equivalent circuit of shielded local oscillator. To restrict the electromagnetic fields to the immediate vicinity of the local oscillator, the oscillator and mixer circuits must be inclosed in a conducting shield which is as continuous as possible. The shield will have to be fastened to the next larger support member at points of equipotential to minimize excitation of the larger surfaces. All power supply leads entering the shielded area must be filtered for radio

frequency disturbances. The oscillator must be designed such that its electromagnetic fields will produce a minimum of current in the shield material. This may be accomplished by (1) the use of a single point ground for the entire oscillator circuit, as illustrated in Figure 3.4.4.1-B, (2) orienting the oscillator coil so that its field induces minimum current in the surrounding metal, and (3) restricting the field due to the oscillator coil by its own individual shield or vestigial shielding such as a shorted turn surrounding the shield. Since the local oscillator drives the mixer or converter vacuum tube, it is necessary to design the bandpass networks associated with the mixer tube for minimum transmission at the oscillator frequencies.



Note: C = 500μμf Low Inductance Type Feed-Through Capacitor First Anti-Resonance Above 300 mc

L = 1μh with C distributed ≤ 0.5μμf

R = All 1/2 Watt With Very Short Leads To Feed-Thru Capacitors

Fig. 3.4.4.1-C Network Arrangement to Prevent Spurious Resonances in Oscillator Range

The best approach in designing an oscillator shield is to enclose the oscillator circuit in a continuously shielded case, using soldered, "water-tight" joints, and even the use of double shielding, if necessary. Where there may be holes and slots in the shield, a difference of potential will exist across them. By proper orientation of the holes and slots, the electrostatic field may adequately be confined. However, this is not true for the electromagnetic field. Magnetic lines of force bulge through any opening and will excite the exterior of the oscillator shield. This, in turn, will excite the main chassis, which will act as a fairly efficient radiator. Holes and slots must be avoided. Where supply leads enter or leave the shielded compartment, they must be prevented from acting as a path of exit for the magnetic lines of force. There must be adequate low pass networks in the power supply leads. In the case of high voltage and automatic gain control leads, a series resistor-shunt

capacitor combination is used. The capacitor must not experience any anti-resonant effects within the tuning range of the oscillator. An excellent arrangement of a network designed to prevent spurious resonances within the oscillator tuning range of ordinary television receivers is shown in Figure 3.4.4.1-C. The use of a few hundred ohms in the output and input of this network helps to prevent resonance occurring in the supply leads exterior to the tuner. L-C networks are usually required in the filament leads to prevent excessive voltage drop. They are designed to prevent any spurious resonances in the tuning range of the oscillator. The capacitors shown in the figure are of the feed-through type.

Coupling between the magnetic field of the local oscillator coil and the shield or chassis must be held to a minimum. This can be accomplished by a high ratio of length to diameter, by a high permeability core which helps to confine the magnetic field, and proper spacing from the chassis. The spacing should not be less than two coil diameters. There is usually a rapid increase in chassis radiation at the high end of the tuning range when using permeability tuners, due to the field extending farther in space when the core is removed from the coil. To properly confine the magnetic field of the coil it is necessary to use both eddy current and permeability shielding. A combination of magnetic and non-ferrous conducting materials will serve the purpose. High conductivity metals are desirable since smaller thicknesses are necessary for a given attenuation of a confined field. The thinner shields also permit the formation of good joints in the shielding. A metal thickness of about ten times the depth of field penetration will produce an attenuation of approximately 86 db in the field intensity. At a minimum frequency of 80 mc the minimum amount of copper required would be 0.003 inches.

The oscillator shield joints must have as large an overlap as possible to prevent leakage. This may be accomplished by screws or spring pressure (multiple contact type) which assures continuity in the shielding. If overlays of copper are used, the shield should be formed with the copper on the inside. Cold rolled steel plated with copper offers adequate attenuation, but requires greater thickness of shield, and is subject to corrosion. If tolerances are of necessity loose, the greatest possible contact is secured in a practical way by the use of a metal textile gasket, as illustrated in Figure 3.4.4.1-D.

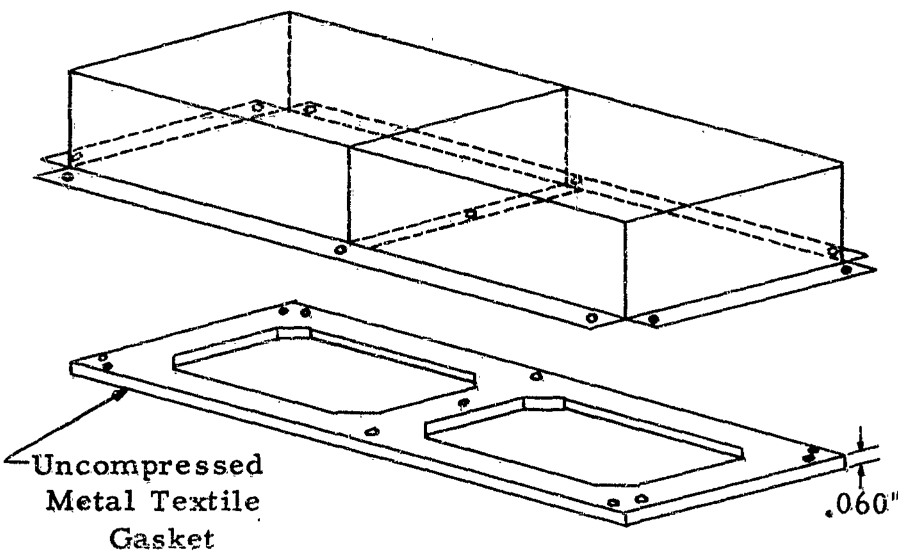


Fig. 3.4.4.1-D Typical Shield Construction

3.4.4.2 ANTENNA RADIATION

Antenna radiation can adequately be suppressed by the measures indicated under Paragraph 3.4.2. These are additional radio-frequency stages, extra tuned-link circuits, proper shielding and filtering, adequate RF by-passing, and the use of short leads and compact layout. A well screened radio frequency amplifier, together with proper shielding, prevents the energy of the oscillator coil from getting into the radio frequency amplifier grid circuit. The shielding processes employed by manufacturers of very high frequency standard signal generators represent some of the best present-day techniques for the purpose of confining local oscillator energy.

However, a well shielded and properly filtered receiver is frequently a source of radio interference due to the local oscillator voltages appearing on the antenna. This generally occurs when all the possible paths for interference to reach the antenna are not taken into consideration during the design of the receiver.

This difficulty cannot be corrected simply by improving the shields or filters; since the local oscillator output is coupled to the antenna by the circuit elements themselves, rather than through improper shielding or lead filtering. An optimum design must be achieved to reduce the effective coupling between the oscillator and antenna terminal through the mixer and RF stages, and still not adversely affect the operating characteristic of the converter stage. In general, additional RF stages are the most effective means available to accomplish this. Refer to Paragraph 3.4.2 for a detailed treatment of the use of additional RF stages in reducing local oscillator radiation.

A converter section of a typical receiver that was producing high level interference signals is considered herein to illustrate the problems and techniques involved in reducing radio interference caused by local oscillator antenna radiation. Figure 3.4.4.2 shows a simplified schematic diagram of the converter section of a typical radio receiver. The paths over which the local oscillator signals travel to reach the high and low band antenna terminals are represented by a series of dots.

The oscillator signal at the antenna terminal may be reduced in several ways. High "Q's" of the signal frequency tank circuit help to attenuate the oscillator signal appearing on the antenna. A limit is set by the increased difficulty of tracking and alignment of the various stages with increase in "Q". This limit was actually reached in the redesign of the preselector in the sample receiver under consideration and alignment difficulties due to the sharpness of the preselector were considerable. The "Q" of the mixer tank should also be increased, particularly in the frequency range where the largest oscillator signal on the antenna occurs. Test results showed that because of the above limitation only minor improvement could be obtained through increased "Q's" in the circuitry.

The oscillator is capacitively coupled to both the low and high band mixer circuits. Reduction of this coupling would certainly reduce the interference appearing at the antenna terminals. However, since this capacitor also couples the oscillator frequency into the mixer stages for superheterodyne receiver operation, excessive reduction of coupling adversely affects the receiver operation.

The following chart shows the effect of varying the coupling capacitor between

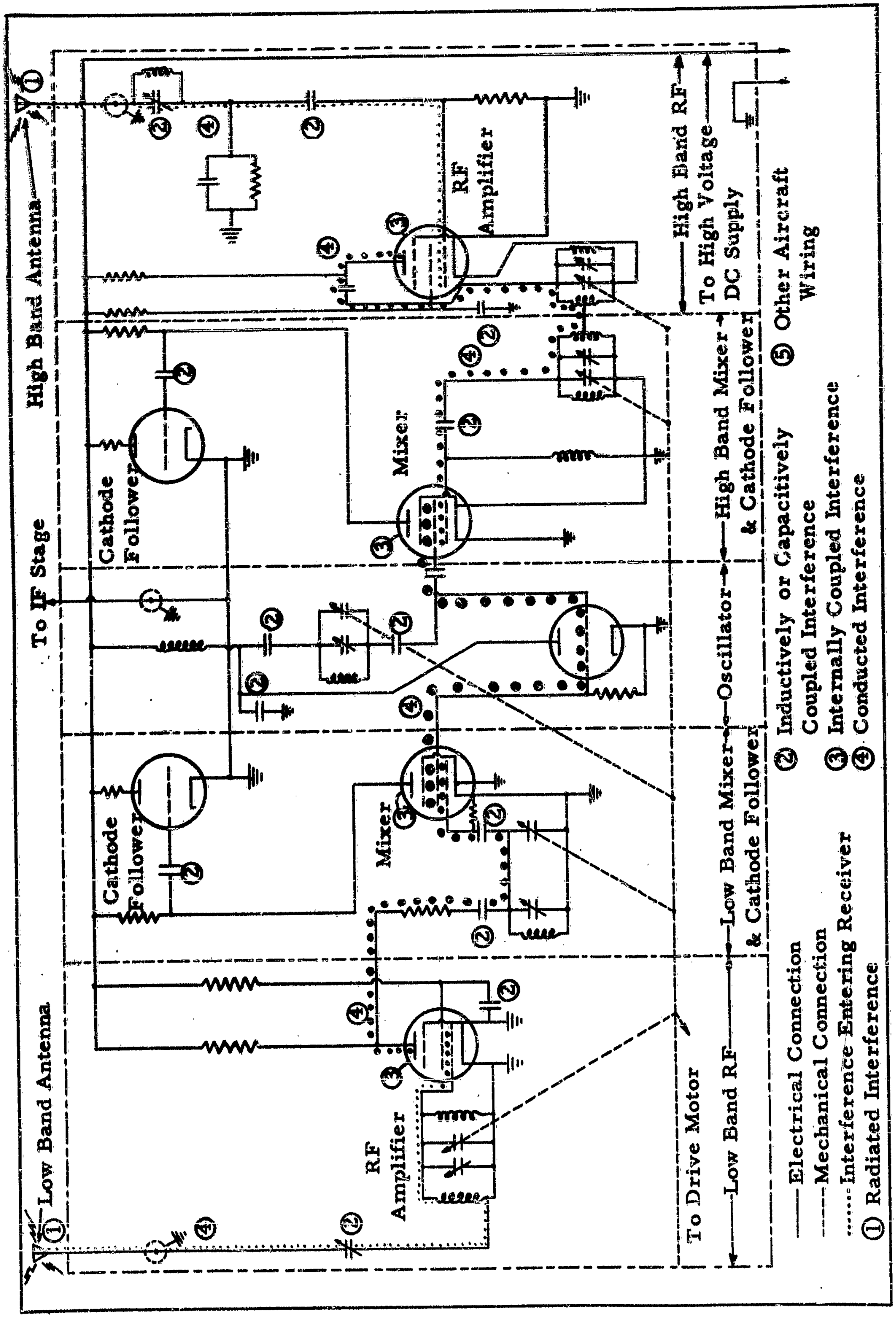


Fig. 3.4.4.2 Path of Local Oscillator Interference Signals in the Converter Section of a Typical Scan Type Receiver

the oscillator and mixer circuits on oscillator radiation and sensitivity in the typical converter section.

Coupling Capacitor	Oscillator Signal	Sensitivity
10.0 μμf	28,000 μv	10 μv
2.5 μμf	8,000 μv	14 μv
1.0 μμf	3,000 μv	25 μv

This clearly demonstrates that considerable oscillator attenuation by this method results in an appreciable loss of sensitivity. For this reason only minor improvement can be achieved through network coupling change.

Oscillation weakening by reduction of plate voltage, or by any other means, is in a class similar to the coupling problem. If the output of the local oscillator is reduced, the interference characteristics are improved at the sacrifice of performance. In fact there is danger of the oscillator failing to oscillate at the low frequency end especially with a weak oscillator tube.

Reference to Figure 3.4.4.2 shows a coupling loop between the oscillator and high band RF stages. It is conceivable that sufficient capacity between the tank inductance and the coupling loop could affect the oscillator coupling. Tests performed on the typical receiver showed that this capacity was sufficient to produce a near-null in voltage at the center of the loop. However, further tests revealed that capacity coupling of the loop to circuit elements resulted in opposing voltages which tend to cancel, and reduce the overall coupling effect. The effect of the total capacity between loop and coil was also investigated by increasing the capacity of one end of the tank coil to ground, thus increasing the voltage between tank coil and loop, also destroying the cancellation mentioned above. Since no appreciable change in oscillator coupling was observed, it was concluded that the total coil to loop capacity, even without the benefit of the cancellation effect, produces negligible interstage coupling.

The above measures apparently must be considered as second order effects only. A device is required that will attenuate the oscillator voltage appearing on the antenna terminals without adversely affecting the sensitivity or other desired characteristics of the converter. Another RF amplifier stage should accomplish this since the sensitivity is improved by the additional stage, and the plate to control grid or screen grid to control grid capacitance coupling is relatively low. The choice of electron tube to be employed is quite important since the capacitance coupling varies with different tube types.

Several experiments were performed to verify the above conclusion. A 10,000 microvolt signal was applied to the plate of a triode and a 7,000 microvolt signal was measured on the grid. A pentode was then substituted for the triode and the test repeated. In this case only 500 microvolts appeared on the control grid. Thus an attenuation of 26 db was obtained with one RF stage using a pentode. A further improvement can be effected by using a pentode mixer with screen grid injection. From the above test results it was concluded that the most effective way to reduce the oscillator signal on the antenna is to add an RF pentode preamplifier stage.

The final design of the converter section crystallized on the use of a pentode preamplifier stage incorporating the above principles. In this model the oscillator signal on the antenna terminal is 30 microvolts maximum at approximately 240 mc. Sensitivity is 12 microvolts where tracked, and image sensitivity is from 6500 to 80,000 microvolts. The following list shows the design considerations that produced the maximum reduction in the oscillator radiation without adversely affecting the overall operation:

(a) pentode RF stage,
(b) screen injection of oscillator signal into mixer,
(c) connecting coupling lead near bottom end of the grounded tank plate circuit coil and near bottom of the grid tank coil of the high band mixer,
(d) shielding between the input and output sides of the pentode socket.

The design problems presented above and the modifications made in the design and development of a typical receiver serve to illustrate the necessity for minimizing interference signals at the source.

These considerations also serve to emphasize the fact that considerable time and expense can be saved by application of sound interference-free design techniques in the original design rather than attempting to correct an interfering component or system once it has been installed.

4. PRECIPITATION STATIC

Precipitation static, as the name implies, is radio interference experienced when the flight path is through precipitation. Flight tests carried out over a period extending back several years have served to tie down some of the principal causes of the interference, although there are some aspects of it which could profitably be explored more fully. The interference is ordinarily intense in the low and medium frequency ranges, and often completely blocks communication or navigation on equipment such as the liaison set, the radio compass, or Loran. VHF equipment is not ordinarily affected, although some few reports of interference to VHF under severe conditions have been received.

Current knowledge of the causes of precipitation static shows that high-speed aircraft with properly designed dragless antennas should be relatively immune to such interference. Attainment of immunity is dependent upon development, specification, and proper application of conducting coatings and upon engineering designs based on a knowledge and proper consideration of the requirements of high corona thresholds. Possible future development of superior methods of controlling aircraft-charging characteristics may eventually ease present design requirements.

4.1 CHARACTERISTICS OF PRECIPITATION STATIC

Precipitation static interference is characteristically broad-band, with a continuous spectrum and produces a very loud "rushing" sound in the output of a receiver, similar to amplified "shot" noise. Since it occurs when the aircraft is on instruments and in need of radio communication and navigation facilities, the interference presents a serious flight hazard.

Flight investigations have established that disruptive discharges in air are responsible for practically all of the interference. These discharges produce steep-fronted impulses of relatively high peak-magnitude, and the interference may be severe when the average current transported by the discharges is quite small, of the order of microamperes. In general, such discharges arise as a result of charging effects produced by impact of the airplane surfaces with precipitation, although they may also occur as a result of electric fields surrounding charged clouds. Impact charging may produce interference as a result of the charging of the whole aircraft or as a result of localized charging confined to the vicinity of the antenna.

4.2 ATMOSPHERIC CONDITIONS FAVORABLE TO PRECIPITATION STATIC

Radio interference of an atmospheric nature was observed long before it became of interest in connection with aeronautics. Radio antennas were observed to collect static charges during snow and dust storms, causing severe interference with the reception of radio signals. Metal windmills on insulated towers, and gasoline trucks on rubber tires, accumulate high potentials whenever they are hit by particles of dry snow, sand, or dust. Aircraft, operated at much higher speeds, experience more rapid charging which results in more serious radio interference.

The major portion of experimental work on precipitation static has been primarily concerned with problems of the charging of the aircraft as a whole, because effects of charge accumulation on the airframe are responsible for most of the static experienced on reciprocating-engine aircraft with external antennas. Increased use of integral antennas has placed more emphasis, in recent years, on problems of localized charging effects, although it is still necessary to give consideration to overall charging characteristics.

4.2.1 DUST, SAND, AND SMOKE

The impact of sand or dust particles against a metallic surface produces, by friction, a charge both on the dust particle and on the metal. If a positive charge is carried away by a sand blast, the metal becomes negatively charged. The same effect would result from the impact of any form of dry powder against a dry surface. The carbon particles which constitute smoke would therefore also be effective in producing a charge. Aircraft flying through the dust storms of deserts and drought-stricken areas have repeatedly become charged in this manner. The denser the smoke or dust cloud and the higher the relative velocity of the wind and the airplane moving through it, the greater the number of impacts per unit area per second, and therefore the more rapid the charging rate.

4.2.2 ICE CRYSTALS, HIGH ALTITUDES, LOW TEMPERATURES

Fine ice crystals, such as are encountered in driving snow at high latitudes or in the fine ice spicules which compose cirrus clouds, are an effective source of precipitation static. Such ice crystals occur only at low temperatures - ten degrees below zero Centigrade, or less. The height of cirrus clouds varies with the season and with latitude, but they do not usually form below 30,000 feet. Ordinary dry snow crystals are produced very readily in cold weather at intermediate altitudes. Flying through dry snow is one of the most common causes of severe precipitation static, a fact which has been demonstrated many times in flights to the Aleutians, and to Iceland and Greenland.

Aircraft flying through snow almost always become negatively charged, and discharges from charged aircraft, whether by corona or otherwise, consist of negative electricity. Laboratory tests in which dry snow has been driven at high velocity against insulated airplane surfaces have also resulted in the production of negative charges on the plane.

4.2.3 FRICTIONAL CHARGING BY IMPACT WITH WATER DROPS

While flight through dry snow or through the fine ice spicules of cirrus and alto-stratus clouds invariably results in precipitation static, flying through wet snow, glaze, or rain usually produces relatively little radio interference. For completeness, however, it is desirable to consider it briefly and to point out that there is no sharp line of demarcation between the various cloud formations, but rather, a gradual shift from one form to the next. Furthermore, in the turbulence above frontal storms, dry snow, wet snow, glaze, and rain may all be present simultaneously at different levels.

4.2.3.1 FLIGHT THROUGH WET SNOW

Wet snow is encountered only in regions where the ambient temperature is near or slightly below the freezing point, and where the relative humidity is high. Frictional charging requires dry surfaces and low relative humidity. The impact between the plane and a snowflake produces energy, $E = mv^2/2$, where m is the mass of the snowflakes and v is the numerical sum of the components of their velocities in the direction of flight. The heat energy thus released melts more snow to increase the magnitude of the liquid film on the surfaces of the airplane. Snowflakes that remain solid after impact are carried away by the airstream together with excess raindrops, taking with them any charges previously left on the metal surfaces. Should the aircraft enter a region filled with wet snow with temperatures of its metallic surfaces below freezing, the snow flakes would form an ice coating over which the impinging snow would slide with only mild charging effects.

4.2.3.2 FLIGHT THROUGH FREEZING RAIN OR GLAZE

Where a frontal condition exists in which warm, moist air has been lifted above a colder air mass, snowflakes falling from higher levels are turned into fine, supercooled rain drops and then descend into the heavier and colder frontal regions below. An airplane, going through a misty and foggy region which separates the warm, upper layers from the cold, frontal air masses near the ground, becomes covered with moisture particles that are small and quite near the freezing point. Such moisture particles seldom are electrically charged to any great extent but easily become attached to all the metal surfaces of the airplane and produce an even more serious hazard by adding a heavy load of ice to all exposed parts.

4.2.3.3 FLIGHT THROUGH RAIN DROPS

Many laboratory experiments have been performed on rain drops falling at various velocities and under a large variety of electrical field conditions. It has been demonstrated that there must be an updraft of more than eight meters per second in a thunderstorm in order to support small hailstones and the larger sized rain drops. It has also been shown that rain drops falling with greater velocity than eight meters per second are torn to pieces by friction with the air through which they fall. Furthermore, these falling drops are charged, some positively, others negatively. The main body of a large drop retains a positive charge while the smaller droplets which have been torn away from it carry negative charges. It follows that the splashing effect of large rain drops against a metal surface moving at high speed must result in the breaking up of the rain drops encountered, the larger drops adhering to the metal structure and the smaller ones carrying negative charges away. Since the airplane is usually charged with negative electricity where precipitation static is present, the positive charges on the larger fragments of the rain drops tend to discharge the airplane and thus decrease the amount of radio interference due to precipitation static.

4.2.4 CHARGES INDUCED BY EXTERNAL FIELDS: THUNDERSTORMS

Precipitation static, quite independent from that resulting from frictional effects, although it may exist coincidentally, is also caused by the passage of the airplane through an electrostatic field such as that existing between two oppositely

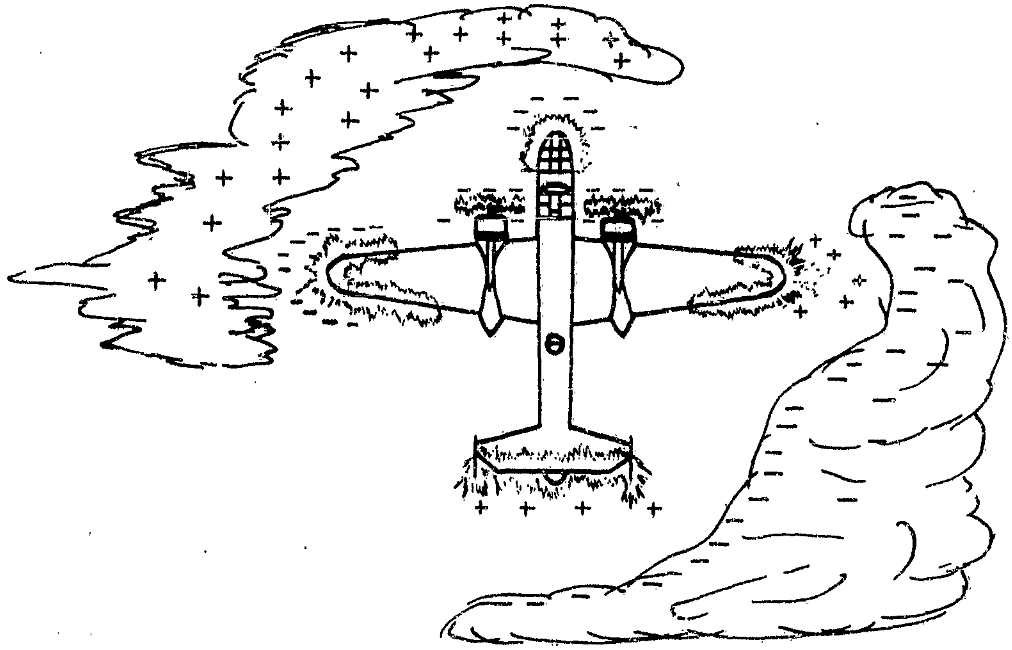


Fig. 4.2.4 Charge Accumulations on an Aircraft in an External Field

charged clouds. The greatest charge per unit area is built up around sharp edges or points such as antenna wires, propeller tips, and the edges of wings, ailerons and rudders as shown in Figure 4.2.4. Maximum corona discharges, which are the source of radio interference, always occur from these sharply curved surfaces, whether produced by frictional impact or by induction from surrounding fields. Therefore, the aircraft designer should avoid sharp edges or points in order to reduce radio interference.

4.2.4.1 EFFECTS OF ALTITUDE AND TEMPERATURE

Charging by induction is limited to those synoptic conditions in which charged clouds produce an electrostatic field. Such fields may be vertical or horizontal, or a combination of the two. They are far from uniform and are constantly varying, especially under thunderstorm conditions.

It is pointed out in Paragraph 4.2.2, that flight at the cirrus cloud level (about 30,000 feet) probably results in severe frictional charging but that this altitude is usually free from electrostatic fields. Descent into the lower levels of non-turbulent clouds of the cirro-stratus, cirro-cumulus (about 25,000 feet), or alto-stratus type (about 12,000 feet) may relieve radio interference resulting from frictional charging as well as from charging by induction. This is particularly true of dense, fog-like clouds since, if flight is through fog or vapor, there is no charging. However, depending somewhat on the height of the cloud above the flight level, discreet precipitation particles may be encountered and charging would still occur. The cirro-stratus level, depending on the latitude and the season, at 12,000 feet is approximately the upper limit for liquid rain-drops, and at about this altitude they occur only together with freezing rain or snow at temperatures near 0°C (32°F). Above this approximate altitude, or freezing isotherm, snow is found which produces static interference due to friction. Below this altitude, icing conditions and induction

charging in electrostatic fields between clouds, or between the clouds and earth, are encountered. In winter this freezing isotherm may be near the ground with frictional charging encountered from the ground level upward.

4.2.4.2 EFFECTS OF TURBULENCE IN THUNDERSTORMS

Turbulence is a prime characteristic of thunderstorms which in turn produce conditions for radio interference in aircraft as a result of induction and frictional effects. Where extensive local heating of air with high humidity results in a steep temperature gradient from the ground to the condensation level, a "heat thunderstorm" may develop. Adiabatic expansion of moisture-laden air in the ascending column causes precipitation with a resulting release of latent heat, which in turn warms the air and increases its velocity of uprush. By such means are produced the rapid convection currents necessary for the electrification of clouds. A "cold-front" storm results when cold, dry air underruns a mass of warm, moist air and lifts it to the condensation level, while a "warm-front" storm is produced when a mass of warm air is pushed up over a comparatively stationary mass of cold air. In each case, the conditions producing turbulence are present.

The typical storm in temperate regions is the cyclonic "low". The low-pressure areas which move slowly from northwest to southeast across the American continent are several hundred miles wide. The general direction of rotation in the Northern Hemisphere is counter-clockwise. In such a general storm area, there is a warm front in the southeastern section and a cold front in the western and northwestern section. During the thunderstorm season, which extends from March to October in the temperate zone, the air in the warm front has originated in the Gulf of Mexico and adjacent land areas and has a high humidity. Its temperature is higher than the temperature of the surface air over which it flows. The cold front, on the other hand, consists of comparatively dry, cold air which has originated in the Canadian Rockies or in the plains toward the east. When such a mass of cold air rushes into the warmer surface air of the plains region, its approach is characterized by a long line of thunderstorms and corresponding turbulence. Such a "line-squall" may have a length of 400 to 500 miles and a height of 6 to 8 miles. In the tropics or near large bodies of water where the ascending air currents have an unusually large moisture content, the total height of the turbulent condition is usually less than 5 miles. In either case, it is obvious that a pilot will not be able to fly around the end of this storm area, rather than pass through its center, unless he happens to be near its periphery. It is also evident that he will seldom be able to fly over the top of such a line-squall without encountering precipitation static of the frictional type. Since the storm conditions are far from uniform along the squall line, he may be able to find a gap of minimum severity through which he may safely pass.

Non-frontal thunderstorms are commonly called "heat thunderstorms" and may be encountered in both tropical and temperate latitudes. Heat thunderstorms usually reach their climax in late afternoon while line-squall storms may arrive at any time during the day or night. Heat thunderstorms are considerably less extensive than the other type and can ordinarily be circumvented without the plane getting seriously off course. Non-frontal storms may, however, develop into general storm areas without the presence of any frontal phenomena. These constitute a special hazard to the pilot on cross-country flight under generally settled weather conditions.

4.2.4.3 MAGNITUDE OF ELECTROSTATIC FIELDS AND THEIR EFFECTS IN THUNDERSTORM AREAS

Laboratory tests show that the potential necessary to produce a spark between polished balls, 10 cm in diameter and 1 cm apart, at 76 cm of mercury pressure in dry air is about 30 kilovolts. For needle points, it is only 12 kilovolts, and is less at high altitudes where the pressure is lower. Field strengths of 400 volts per cm have been frequently encountered in test flights, and a field of 3400 volts per cm has been recorded just before the airplane was struck by lightning. At 1000 volts per cm, a lightning stroke 3-miles long from cloud to cloud, or from cloud to ground, requires an initial difference of potential of about 500 million volts.

In frictional charging through snow or freezing rain, or under dust-storm conditions, the same sign of charge, usually negative, is present over the entire plane and the density of charge at any given point on the aircraft remains essentially constant for comparatively long periods of flight. On the other hand, induced negative charges on the plane are found nearest the positively charged clouds, while the positive charges are found nearest the negatively charged clouds. In thunderstorm fields, both the sign and the potential gradient of the charges which cause induction vary rapidly with both time and the position of the plane with respect to the active portion of the storm. Resulting currents of an order of magnitude in the milliampere range surge back and forth from wing tip to wing tip and from nose to tail, all capable of producing extreme radio interference due to spark-over at poorly bonded joints.

The electrical capacity, in flight, of even a large airplane is quite small; for example, it is about 780 μμf for a B-17. Therefore, a current of 100 microamperes flowing into it would charge a B-17 at the rate of 128 kilovolts per second. The sharp edges and points would therefore quickl reach the corona point when in the presence of strong electrostatic fields and cause severe radio interference.

4.2.4.4 LIGHTNING DISCHARGES TO AIRCRAFT

Although the all-metal structure of aircraft affords protection to the crew against lightning discharges, a direct discharge across a metallic conductor, such as an airplane, still presents three problems. Radio interference is at a maximum but usually of short duration due to the stroke itself. The after-effects of the stroke may actually be more serious than the radio interference itself. The current in a lightning stroke may be as much as 100,000 amperes. While the surface and framework of the plane can carry such a current without danger, even a small fraction of such a current, if it gains access to radio receivers, power supplies, and other electrical equipment, can easily fuse wires and put control apparatus out of operation. It may also start fires by jumping across poorly bonded sections of the plane, especially in the vicinity of gasoline leaks which may be present. When flying at night, the intense illumination resulting from a lightning stroke may blind the pilot so that he cannot see his instruments for several minutes. This blinding effect can be minimized by keeping the cockpit light on full when in the vicinity of electrostatic fields. While there is little danger of electrical shock to persons inside the closed cockpit, there may, nevertheless, be resultant nervous reactions which render the pilot more or less incapacitated for the performance of his duties.

The distribution of electrical charges in a thunder cloud is far from uniform, and authorities are not fully agreed on any general pattern. Whatever the distribution of electrical charges within the cloud, the potential difference between points at a distance equal to the wing span of an airplane undergoes rapid changes in sign and magnitude according to the part of the cloud in which the airplane is flying at any given moment. If it is in the region between maximum negative and maximum positive charge concentrations, the potential gradient is high. On the other hand, where the distribution of charges is of nearly the same concentration and of uniform sign, the maximum currents and potentials connected with a stroke through the plane cannot be very large. The probable minimum risk is taken by flying as high as possible above the active vortex of the storm or by flying to one side or the other of that vortex. When confronted with an extended line-squall storm with towering cumulus clouds that reach above the ice-spicule level, the pilot may be justified in flying under the cloud in the raid-area between the cloud and the earth. This involves three hazards, the icing and induction-charging dangers discussed in Paragraph 2.1.4.1, the possibility of intercepting a lightning stroke in the high potential gradient between the cloud and the earth, and the danger of encountering mountains or other obstructions.

4.3 METHODS FOR REDUCING INTERFERENCE FROM PRECIPITATION STATIC

Increased speed and higher altitude flying, the result of improved design and the availability of more powerful engines, both dictated by military necessity, brought ever-increasing losses due to the precipitation-static type of radio interference. The Military Services, aircraft manufacturers, and commercial flight operators have made attempts to remove this hazard to aerial navigation. From numerous investigations, in flight and in the laboratory, there resulted a series of programs for "cleaning-up" airplane structure and design. These included the elimination of sharp-edged or pointed protuberances, the effect of various coatings on the metal surface, and the bonding of separated parts of the aircraft structure. In addition, much research and development has been done to make available dielectrically insulated antenna wire and fittings as well as devices for discharging static charges harmlessly from the aircraft while in flight.

4.3.1 OPERATION OF AIRCRAFT

Pilots have frequently observed that the severity of precipitation interference is a function of the speed. Using as a working equation, charging current equals a coefficient times the speed to the n^{th} power, it was found that the average value of n is 3. Thus the radio interference goes up approximately as the cube of the speed. Reduction of airspeed from 400 miles per hour to 200 miles per hour would, therefore, reduce the charging current, and hence the corona discharge, by a factor of eight as shown in Figure 4.3.1-A.

The variation of the coefficient with temperature is not linear. It is a maximum between -7° C and -9° C, while the exponent, n, reaches its peak value at about -7° C, as is shown by the curve in Figure 4.3.1-B. The severity of precipitation interference may therefore be reduced both by decreasing speed and by descending to lower levels where the temperature is higher. The second alternative is, of course, always accompanied by the dangers caused by icing.

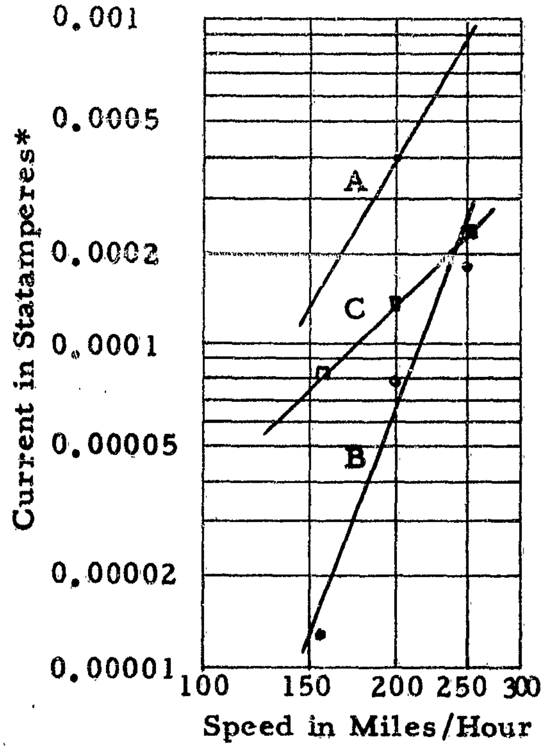


Fig. 4.3.1-A Charging Current as a Function of Speed & Temperature
Curve A - Temp. -83°C Slope n=3.3
Curve B - Temp. -67°C Slope n=4.7
Curve C - Temp. -50°C Slope n=2.1

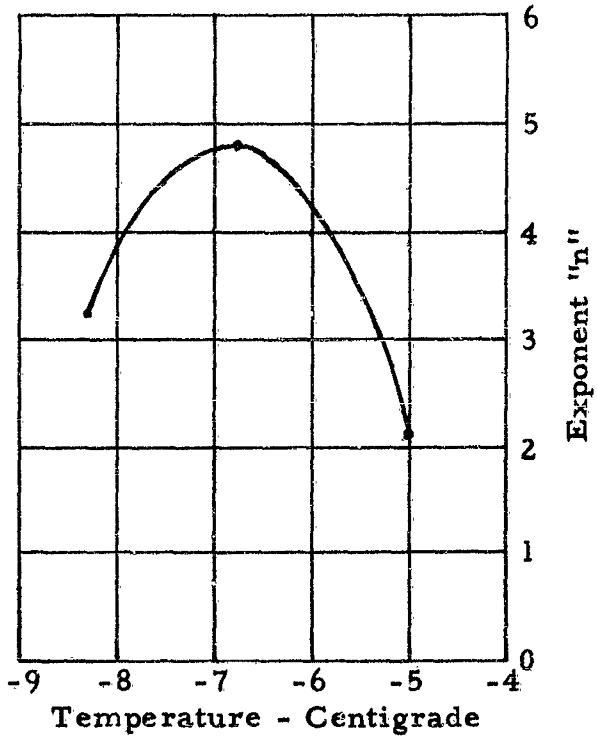


Fig. 4.3.1-B Temperature Effect on Exponent in the Equation: Charging Current = Coefficient $(Speed)^n$. (Nose-piece coated with aluminum.)

4.3.2 EFFECTS OF COATINGS ON METAL SURFACES

Both as a means to protect a shiny aluminum surface against the elements and to reduce visibility to anti-aircraft and other enemy fire, airplanes have customarily been covered with various varnishes and camouflage paints. It has been shown that the charging rate of painted surfaces is several times as great as that for clean aluminum surfaces. Surfaces of clean aluminum and those coated with aircraft wax produce negative charging, but coatings of TiO_2 and colloidal silica give positive charges when bombarded with driving snow. The results of these and other tests show that the paints in common use on aircraft surfaces consistently give negative charges to the airplane surface under precipitation charging conditions.

The fact that coatings with positive charging coefficients are available has led to attempts to produce a chargeless airplane by using equivalent amounts of positive and negative materials on different sections of the plane. Experience proved, however, that contamination of positive surfaces, caused by handling of the plane by the ground crew, resulted in a reversal of sign for the charging coefficient. The best possible surface for minimum precipitation static is clean, smooth aluminum. The charging coefficient, K, for various coatings as a function of temperature is shown

*One Statampere (electrostatic cgs units) is equivalent to 3.33560×10^{-10} amperes (absolute).

in Figure 4.3.2. The coefficient, K, is defined by the equation, $K = I/W$, where I is the current, in electrostatic units per square meter of surface exposed to a blast of snow, and W the weight of snow which strikes this unit surface in unit time.

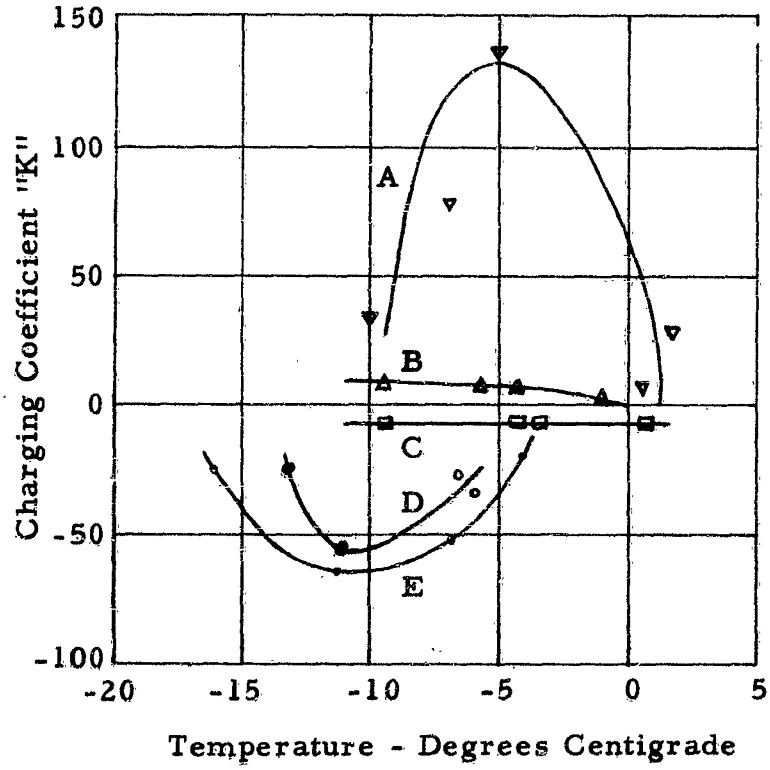


Fig. 4.3.2 - Charging Coefficient as a Function of Temperature For:
A--TiO_2(Anatase Form), Thin Film; Sol No. 155;
B--Colloidal Silica No. 155 in Cellulose Nitrate;
C--Bare 52-S Alumini
D--Aircraft Wax on 52-S Aluminum;
E--Amphibious Transport Paint.

4.3.3 DESIGN OF AIRCRAFT STRUCTURES

From the standpoint of freedom from interference from precipitation static, the ideal shape for an airplane would be a smooth sphere. This being functionally impossible, there remain, as a means of a practical solution to the problem, attempts to reduce the curvature of sharp-edged and pointed structures without interfering with their mechanical operation. This includes rivet-heads, projecting edges of sheet-metal, exposed surfaces of Pitot tubes and thermometers, all of which have received attention in recent aircraft design. Since corona discharge begins at the point of sharpest curvature, it is necessary to reduce the curvature of all critical areas to the same value, so far as possible, in order to keep the potential of the whole aircraft uniformly high and to prevent breakdown below the operating voltage of static dischargers. In some cases, covering the sharp edges with plastic insulation is effective, unless it interferes with normal air-stream flow.

The quantity of electricity, Q, on any metallic body is equal to the product of its electrical capacity, C, multiplied by the potential, V, applied to it, thus $Q = CV$.

The electrical capacity of an insulated sphere in electrostatic units is equal to its radius, r, in centimeters, that is, $C = r$. Electrical charges distribute themselves equally over the entire surface so that $V = Q/r$, but if the sphere is surrounded by a medium of dielectric constant, k, then $V = Q/kr$. In the consideration of aircraft, k is approximately one for air.

The surface density of charge, s, equals the total charge divided by the area of the sphere; therefore,

$$s = \frac{Q}{4\pi r^2} = \frac{V r}{4\pi r^2} = \frac{V}{4\pi r} \qquad (4\text{-}1)$$

For any given potential, therefore, the charge density varies inversely as the curvature, $1/r$. It approaches zero for a plane surface and becomes infinite at the tip of a sharp-pointed needle.

The atmosphere always contains considerable numbers of ions produced by ultra-violet light, cosmic rays, radio-active substances, engine exhausts, etc. Such ions are accelerated towards or away from intense fields, according to their signs, whether positive or negative, and may gain sufficiently high velocities to ionize more molecules by collision and thus initiate a corona discharge with its resulting static interference. Corona discharge generally begins at about 1/2 to 2/3 the potential required for a disruptive spark, although, depending on the geometry involved, it has been observed at 1/10 this potential.

Laboratory experiments have shown that the break-down potential between polished spheres, 1 inch in diameter, is approximately three times as great as for needle-points. One hundred kilovolts can bridge a six-inch gap between needle-points in dry air. Exact break-down voltages have been shown to depend on materials used, temperature, and pressure.

The break-down potential from a metal surface to the surrounding air depends on the density of the air, being a function of the mean free path of the molecules. For aircraft at high altitude the pressure decreases with the altitude while the temperature falls at the same time, increasing the air density. While these two factors affect the air density in opposite directions, the relations are not linear and the pressure effect predominates. At 5000-feet elevation, the break-down potential is about eight-tenths (0.8) and, at 10,000 feet, sixty-seven hundredths (0.67) of that at sea level.

Because of the camber, or curvature, of the upper surface of an airplane wing, the air going over the top of the wing surface must travel farther, and hence have greater relative velocity, than the air passing the comparatively straight under surface. According to the well-known theorem of Bernouilli, the greater the speed of the air over a surface, the less the pressure. Thus is produced the important lifting effect which keeps the plane in the air but incidentally increases the tendency to go into corona over these areas of low pressure. Because of the pitch and relatively high speed of the propeller tips, they are likewise surrounded by a region of reduced pressure as they plough through the ambient snow and ice particles. Consequently, the propeller tips tend to burst into corona almost as soon as bare wire antennas and other sharp metal points. Obviously, both of these Bernouilli effects are inherent and essential to the functioning of the plane. The change in pressure on the wing surface

due to camber is not serious because of its large radius of curvature. The only remedy for the propellers is to bond them effectively to the fuselage to keep their potential relative to the airframe as low as possible, and to keep their surfaces free from offending paints and oils.

4.3.4 IMPROVED ANTENNA WIRE AND INSULATORS

Before the recent advent of antenna wire and fittings of improved design most of the interference to radio reception originated in the antenna and its auxiliary parts. The antenna is of necessity located in an exposed position of high potential. The fine wire of which it was made, to reduce wind-drag, had a very small radius of curvature and the connections to inexpensive strain insulators were crudely fashioned and had sharp-ended, projecting points.

Precipitation static on aircraft employing external wire antennas is usually a result of corona discharge from the antennas. Charging of the whole aircraft is responsible for the corona. This situation has been effectively dealt with by raising the corona threshold of the antennas and providing a discharge path for charges to leave the airframe without creating radio interference. The corona threshold of the antenna may be raised by increasing its radius of curvature or by coating it with insulating material of very high dielectric strength. The Air Force and the Navy now use an anti-corona antenna wire, consisting of a 50-mil diameter, #16 copper-weld, conductor coated to an outer diameter of 183 mils with polyethylene, and the so-called "wick" dischargers on reciprocating-engine transports and bombers.

Antenna fittings currently used with the wire employ smooth metal surfaces having large radii of curvature for increased corona threshold. Figure 4.3.4-A shows the components of antenna assembly AS-315/A.

New fittings are now in production development which will provide complete external insulation for wire antennas. Figure 4.3.4-B shows the components of these new type fittings and antenna masts. Figure 4.3.4-C shows some of the fittings assembled in greater detail. Mechanical limitations make it impractical to apply this type of equipment to high-speed aircraft.

The transmitting and receiving antennas on aircraft are necessarily mounted in such a manner that they are badly exposed to the sources of precipitation static. The relation between the speed of an aircraft and the severity of precipitation static was discussed in Paragraph 4.3.1. The great advances in aircraft speed accomplished in recent years have necessitated a multitude of improvements in aircraft design, one of the most important of which is the adoption of greatly improved antennas, antenna insulators, and masts.

Field and laboratory tests have proved beyond doubt that the corona can be suppressed by using a wire covered with a polyethylene insulation and supplied with fittings which protect the ends of the wires from exposure by layers of insulating material. It was found that a #16 copper-weld conductor with a diameter of 50 mils, insulated by a sheath of polyethylene to bring the total outside diameter to 183 mils was a great improvement over the bare wires previously used. Insulators with rounded corners that could be covered with thick layers of tape were used for connections but it was still found that these fittings were the most likely place for the

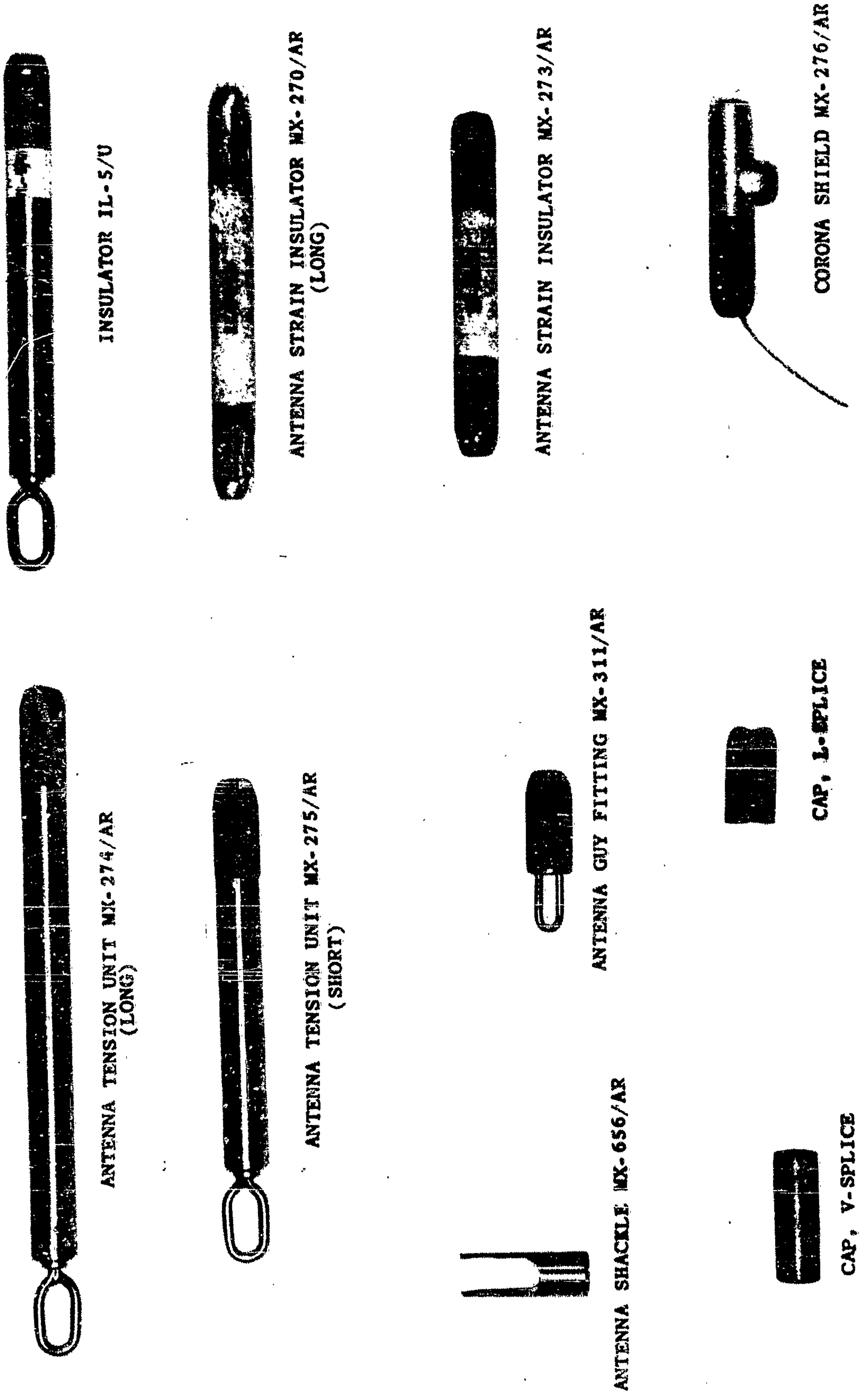


Fig. 4.3.4-A Components of Antenna Assembly AS-315/A

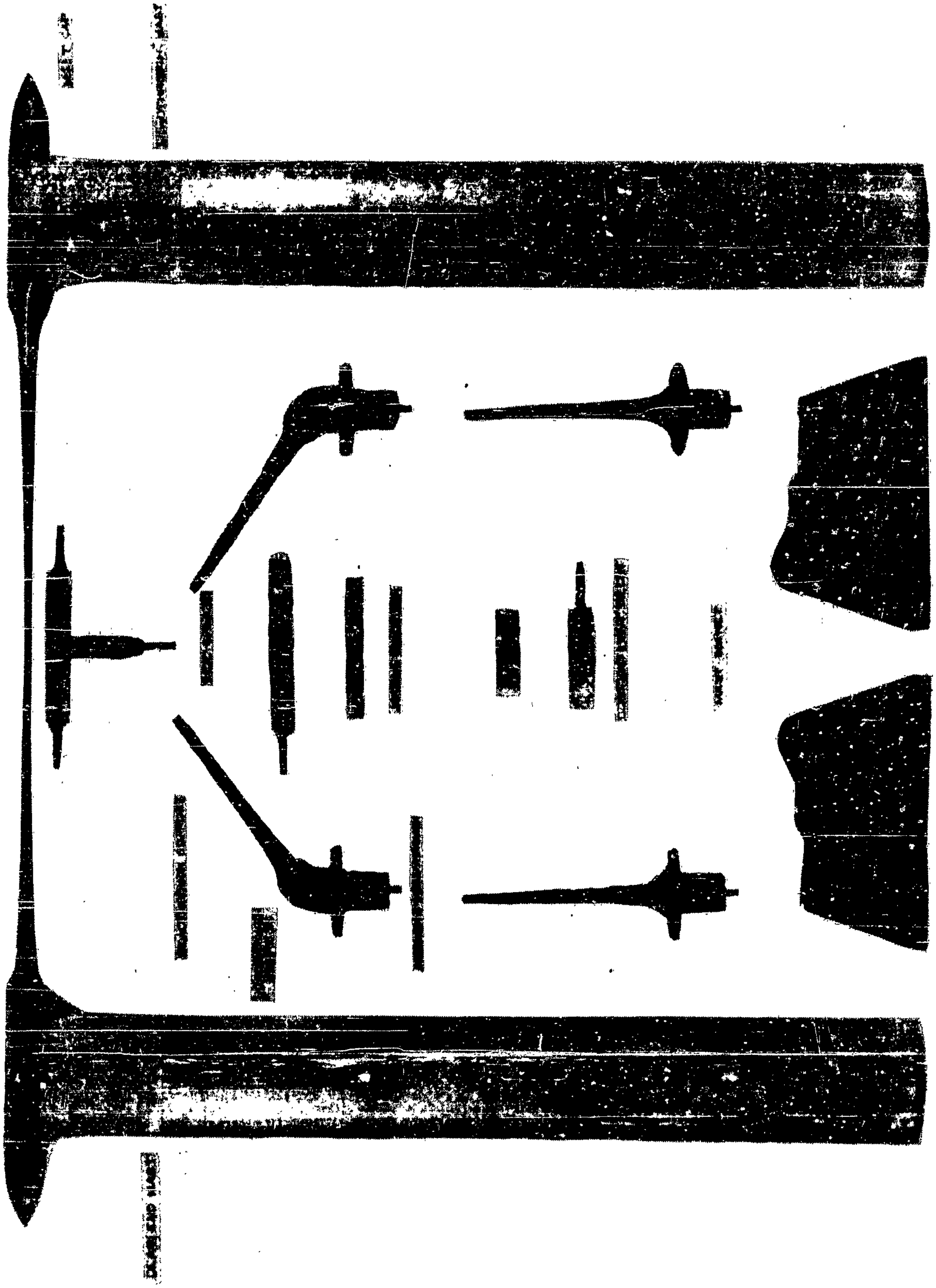


Fig. 4.3.4-B New Type Completely Insulated Antenna Fittings and Masts

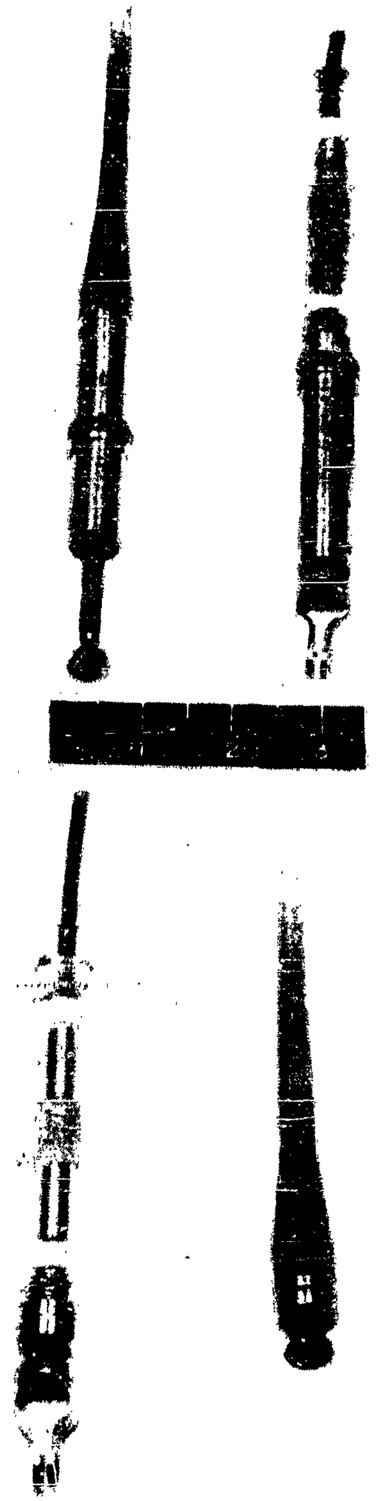


Fig. 4.3.4-C Anti-Precipitation Static Antenna Hardware Assembled

incidence of corona. Difficulty was also encountered from scratches and nicks in the ends of the wires. The new type of antenna hardware includes silicone rubber sleeves, polyethylene thrust washers and molded lucite caps. Instead of coiling the wire around an insulator and twisting its end around a lead-in, a new fitting has been devised into which the end of the wire is forced and clamped between jaws which hold it firmly without twisting.

The new type of strain insulator shown in Figure 4.3.4-B & C contains two wire-holding chucks separated by an RF insulator and encased in a molded sheath which contains a sealing cavity and is closed at each end with a threaded cap. The Tee-Splice also shown contains three wire-holding chucks fastened to a metal tee spacer for mechanical and electrical connection. The insulation and sealing arrangements at the ends are similar to that of the strain insulator. The Dead-End Mast shown in Figure 4.3.4-B consists of a streamlined plastic (Fibreglass) mast 24 inches long, fitted with an adjustable insert at the upper end for protecting and shielding the outer end of the antenna. The insert contains a wire-holding chuck encased in a molded covering of insulating material. The inside of the masthead is threaded to fit the plastic nut on each end of the insert, thus providing a small amount of "take up".

The Lead-Through Mast illustrated in Figure 4.3.4-B has the same general construction as the Dead-End Mast. No insert is used, but a 1/4-inch hole is molded into the center of the mast to permit the antenna wire to pass through so that it may terminate on the inside of the fuselage of the plane. The Mast Socket also shown consists of a pair of molded brackets designed to mount and support either type of mast on the skin of the aircraft.

4.3.5 LOOP TYPE ANTENNAS

The advantages of loop antennas are that they can be electrostatically shielded against high impedance fields, that is, fields for which the ratio of electric to magnetic components is large, they can be brought closer to the surface of the aircraft and can be designed to avoid sharp points or small radii of curvature.

The regulation flat-top antenna are, of necessity, mounted as high as possible in order to intercept the required amount of energy from the desired signal. This is where the electrical fields caused by precipitation static discharges produce strong surges into the receiving equipment. A loop antenna, on the other hand, by making use of a large number of turns in the loop, acquires a larger induced voltage at a lower elevation above the aircraft.

For the condition that the circumference or the perimeter of the loop is small as compared to the wavelength, the induced voltage is given by the equation

$$V = \frac{2\pi}{\lambda} ANE \cos\theta \tag{4-2}$$

where V is the induced voltage, A the area of the loop in square meters, λ the wavelength in meters, E the field strength in volts/meter, N the number of turns, and θ is the angle between the incoming wave and the plane of the loop. The maximum value for V is obtained when $\cos\theta = 1$. The effective length of the antenna, L_e, in

meters, is defined as V/E and has the value

$$L_e = \frac{2\pi A N}{\lambda} \tag{4-3}$$

for the condition $\cos \theta = 1$. The current which flows through a loop of impedance Z is

$$I_c = \frac{V}{Z} \tag{4-4}$$

The reactive part, X_L, of the impedance Z is always inductive for a small loop. If the loop itself is designed so that the inductive reactance is equal to a capacitive reactance which is either designed into the loop or supplied externally, then the current flowing in the circuit would be a maximum and limited only by the effective resistance. In this case,

$$I_c = \frac{V}{R_e} = \frac{V}{X_L} Q \tag{4-5}$$

where Q is defined as the ratio of the reactance, X_L, to the effective resistance, R_e. It is seen, therefore, that the induced current is proportional to the "Q" of the loop.

In using the loop as a radio compass, for example RADIO COMPASS AN/ARN-7 used as a homing device, the width of the loop null is a function only of the radio interference present. As the null point is reached, the automatic volume control in the receiving equipment increases the amplification, and amplifies any interference in the circuit together with the signal. If the receiver is entirely free from interference, the null point may be obtained very accurately, at least to within one or two degrees. One of the disadvantages of the high impedance loop lies in the necessity for physically locating it at such a distance from the receiver that the capacitance of the connecting cable necessitates other changes in the circuit which consequently decrease the effective height of the loop and also reduce its induced voltage. Present practice employs a low impedance loop which is connected to the receiving set through a special transformer and coaxial cable. Either type of loop is enclosed by a metallic shield connected to the airplane structure. When the shielded loop is put in a position parallel to the line of flight, in which case the azimuth angle of the compass needle is at 90°, the loop antenna gives readable signals long after signals from open antennas are lost in crashing noises. In this 90° azimuth position, the loop intercepts the maximum signal and minimum interference. With the local disturbances held to a minimum, the signal voltages can usually be amplified to the point where they are readable above the irreducible interference background.

4.3.6 TRAILING WIRE DISCHARGERS

Trailing wire dischargers were employed several years before Static Dischargers AN/ASA-3 became available. This older type consisted of a fine steel wire connected through a resistance of 100,000 ohms or more to a high potential point at the rear end of the fuselage. Such a wire in actual use was 15 to 30 feet long. Because of the small diameter of the bare wire, there was a tendency for it to go into corona before any other part of the airplane. Corona discharge at the end of the trailing

antenna would often couple back to the receiving antenna and appear as another source of interference. A large resistance in the circuit was used to prevent oscillatory discharges; however, it was found that it was of great importance to have the resistance distributed rather than lumped in one place. In its most useful form, such a trailing antenna is put into a small shield from which it can be released by the pilot through an electromagnetic device operated from the cockpit. A cup attached to the outer end of the wire holds it in the windstream. The use of this device has frequently enabled a pilot to reduce the precipitation interference enough to get range signals through; however, it does add additional hazards so far as lightning strokes are concerned and is falling into disuse. Both commercial airlines and the military have experienced a high incidence of lightning strikes to trailing dischargers.

4.3.7 FLUSH ANTENNAS, CANOPIES, AND RADOMES

Integral antennas may be divided into two classes for consideration of precipitation-static effects. One class is comprised of antennas which utilize portions of the external structure, such as wing and tail-cap antennas; the second is comprised of antennas housed within insulating material.

One problem which is important for integral antennas of the first class is their corona threshold. Flight tests have shown that although jet engines are in themselves fairly efficient dischargers they do not completely compensate for higher rates of impact charging produced by higher speeds. Net charge accumulated on the airframe raises the potential of the aircraft to a point where corona discharge from extremities may occur. Since integral antennas are often located at such points, corona discharge is still a problem. There are three methods of dealing with this situation. The first is to obtain more efficient dischargers, so that the airplane potential remains low. The second is to shape and locate the antennas, if possible, in such a way that they have a high corona threshold. The third method is to control antenna discharges to greatly reduce the magnitude of radio-frequency interference accompanying the corona.

The most fruitful approach at present is that of designing the antenna for high corona thresholds so that large amounts of charge can accumulate on the airframe without causing interference. It seems doubtful that much consideration has been given to this factor in the past, and investigation of some of the aspects of the problem would be very worthwhile.

In general, some of the same factors which provide good efficiencies on low-frequency antennas tend to lower the corona threshold; that is, the antenna may function best if it is located at points where electrostatic field intensification is high. The problem of computing electrostatic field intensities at various points on an aircraft is tedious and difficult. Some measurements of electric field at different points on a typical bomber have been made in flight, and these measurements may serve to indicate the orders of magnitude of some of the ratios. Figure 4.3.7-A shows the points at which measurements were taken, and the ratio of the field at these points to a point near the center of gravity.

Some flight data obtained from corona threshold measurements on dischargers showed that the field intensity at the tips of the horizontal and vertical stabilizers is comparable to the field intensity at the wing tips. Any designs utilizing aircraft

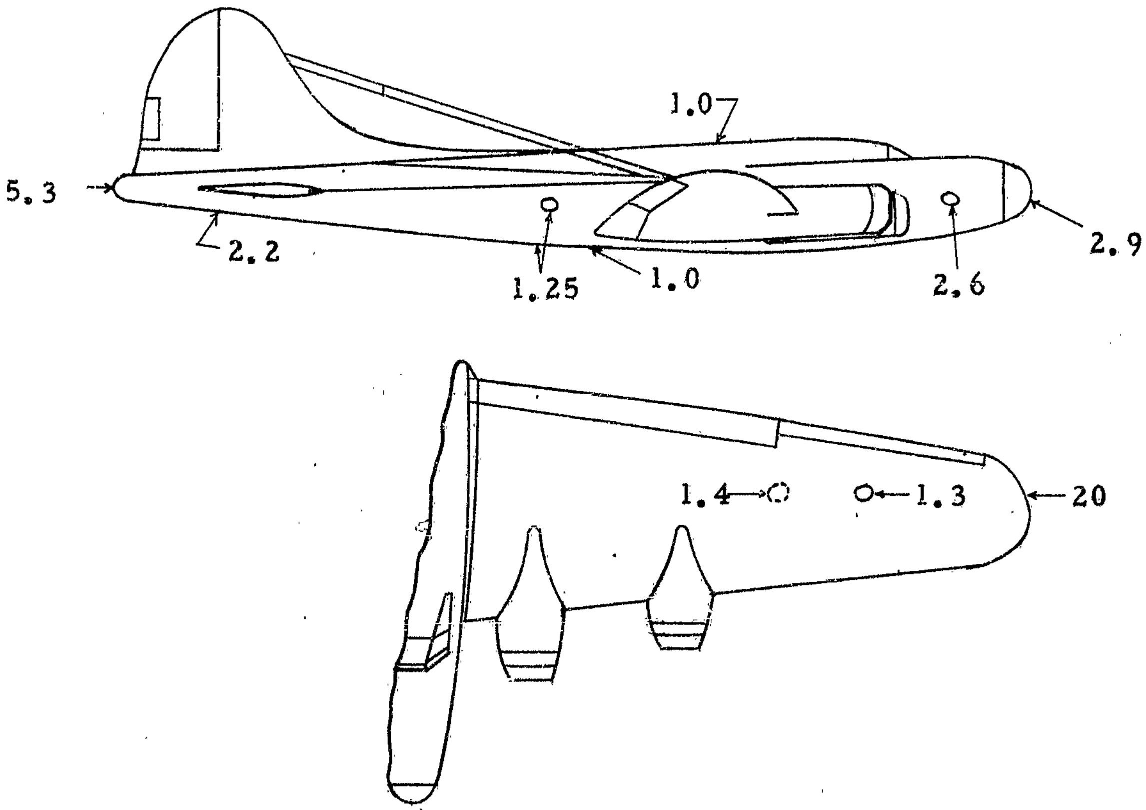


Fig. 4.3.7-A Relative Measurements of the Electric Field as Measured on the Surface of a Typical Bomber in Flight

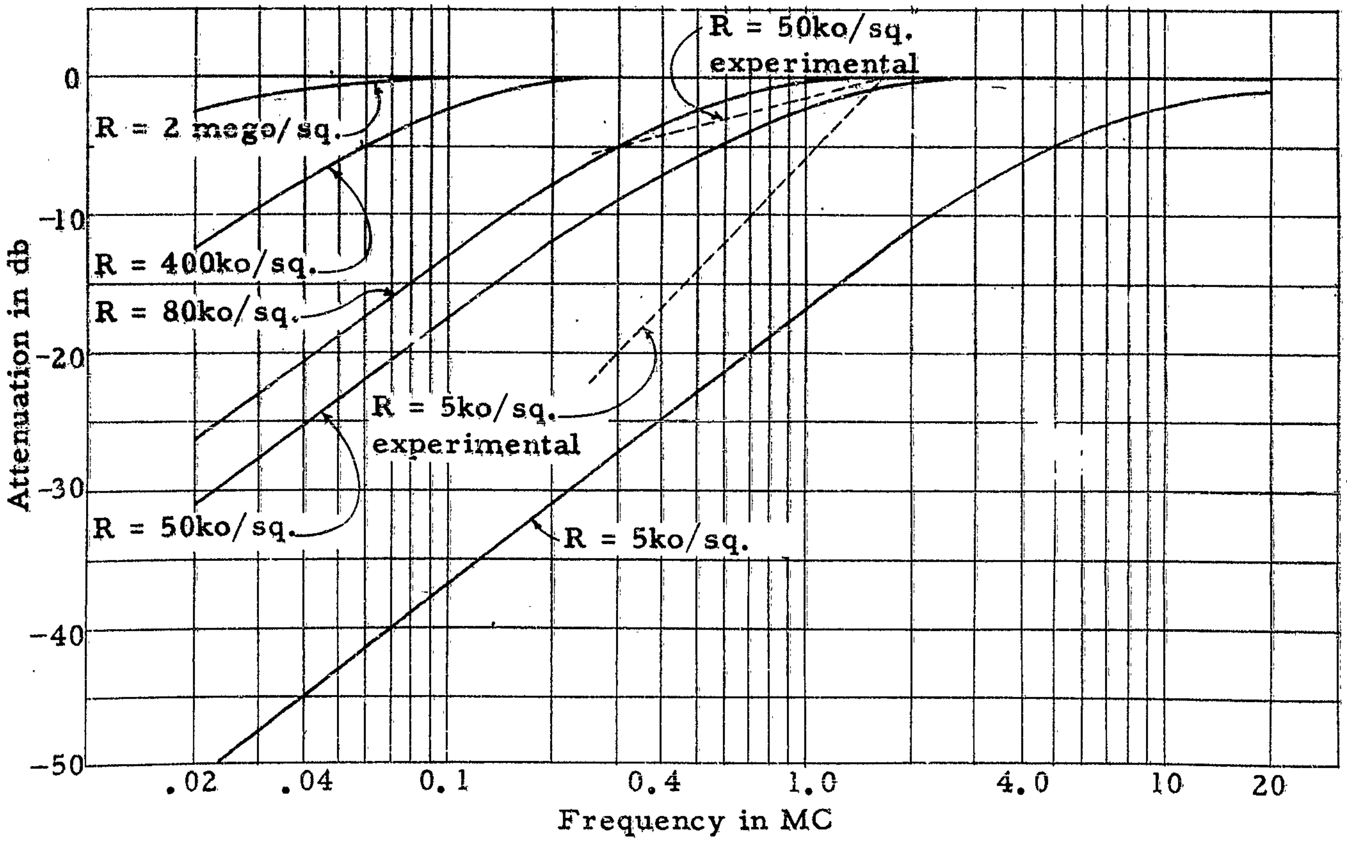


Fig. 4.3.7-B Attenuating Effect of Shield Coatings of Various Surface Resistivities

extremities as antennas should be carefully considered from the standpoint of probable corona threshold, because high values of electric field are obtained at these points with comparatively small amounts of charge residing on the airframe. The best solution would be to avoid placing antennas on the extremities, if that can be done without affecting performance. If performance requirements dictate placement on wing or tail extremities, every effort should be made to maintain large radii of curvature of the conducting surfaces.

A comparatively large number of integral antenna designs utilize some form of housing constructed of insulating material, with antenna conductors close to the inner surface of the housing. If there is no conducting path provided across the outer surface of the housing, it is possible for bound charges to build up on the insulating material as a result of impact charging. When the field gradient between charges or between charge and airframe becomes high, disruptive discharges occur. This "streamering" can produce interference with characteristics similar to those of antenna corona, with the exception that the number of bursts per second may not be as high. This type of interference may be very intense with small charging currents, which results in disruption of reception even in very light precipitation. Radio-compass sense antennas installed underneath plexiglass canopies have proven very susceptible to this type of interference. Plexiglass has excellent insulating properties so that very small charging currents to the plexiglass surface develop high voltage and consequent streamering. Some flight measurements made on the nose canopy of a typical bomber showed 10,000 microvolts of interference developed on a test antenna mounted on the inner surface when the charging current to the outer surface was considerably less than one microampere per square foot. The interference measurements were taken at 300 kilocycles with an effective bandwidth of about eight kilocycles. At the same time the interference was measured on the test antennas, the radio-compass, which uses a sense antenna mounted on the underside of the pilot's canopy, was inoperative on range signals. On this airplane it has also been observed that the interference from the compass loop, which is mounted at the rear of the pilot's canopy, becomes very large at slightly higher charging currents.

An additional undesirable effect can occur from charge accumulation on insulating surfaces. Electric fields are produced on the inside of the housing which may reach intensities sufficient to cause corona from the conducting structure of the antenna. Some measurements of corona current of this type have been made on a typical bomber antenna. The corona is usually intermittent, but in conditions where high values of charging current are present it can contribute significantly to the interference.

Interference-producing effects on insulating structures can be greatly diminished by providing suitable conduction paths over the outer surface. Since the currents involved are quite small, a high order of conduction is not required. If the surface is coated with a thin layer of conducting material, the lower limit of conductivity is set by the maximum voltage difference which is permitted between the airframe and the surface of the housing and the upper limit is set by shielding effects on radio signals. A limited investigation has shown that, in general, higher conductivity is permissible at higher frequencies, and that surface resistivities ranging from three to ten megohms per square are satisfactory for most applications. Calculated and measured attenuation effects at low and medium frequencies are shown in Figure 4.3.7-B. It can be noted that increased losses occur at lower frequencies for a given coating resistivity.

Because pilot's canopies are often used as antenna housings, a need exists for a durable conductive coating which can be applied to plexiglass without degrading the optical properties. The Materials Laboratory at Wright Field is currently engaged in a program to develop a suitable coating. This group experienced difficulty in obtaining a coating in the proper conductivity range. As an interim solution, the application of conducting strips to canopies was considered and tried, but not found completely satisfactory. A promising coating has now been developed and tried with success. Its resistivity is 1-10 megohms per square and has good erosion resistance.

Radio-compass operation in many of the presently used fighters and bombers has been severely limited by precipitation static. These installations utilize sense antennas mounted under plexiglass canopies, which is often the most practical place to install them. The problem is currently a serious one with the military aircraft and manufacturers are currently engaged in running some tests of conducting strips applied to the canopy. These tests are being made with rather high conductivity strips which are probably not optimum for the purpose. Manufacturers have reported that the results of one flight seems to show that the interference was decreased to a point which permitted the radio-compass to operate although reception was still quite unsatisfactory.

The problem of interference coupled from corona or streamering regions to antennas at some distance away has not been given much consideration up to the present time. Coupled interference is more serious for the case of weak-signal reception; scanty observations seem to show that it usually is not a severe problem with moderate signal strengths. Capacitive antennas are apparently more susceptible to interference pick-up from points in corona than loop antennas. For example, corona points on the wingtip of a typical bomber coupled interference to the top-wire antennas. Flight and laboratory tests made with compass loop antennas showed that corona points more than four feet away from the loops did not couple in much interference unless the loops were oriented for maximum pick-up in the direction of the point. Interference coupled to wire antennas was greatly reduced or eliminated if the corona point was screened from the wire by portions of the fuselage. A point worth stressing is that coupled interference may prove troublesome if the more usual types of interference are eliminated.

4.3.8 STATIC DISCHARGER AN/ASA-3

Experiments have proved that charges can escape from sharp points on an airplane without producing radio interference. This helps to keep the aircraft at a low potential with respect to the surrounding atmosphere and thus minimizes the possibility of corona discharges. Steel needle-points have a tendency to become corroded so that they are no longer sharp points. Fine cotton fibers have been found to be effective and practical when formed into a so-called wick, as in the AN/ASA-3 Static Discharger. Each fiber serves as a separate point from which charges can escape. The cotton is treated with a compound which reduces its resistance but keeps it high enough so that the current escaping from each fiber is very small and cannot become oscillatory because of the large resistance in the discharge circuit. The dischargers are mounted at points of maximum electric intensity and removed as far as possible from antenna leads to reduce coupling effects. They are not completely effective but are capable of very greatly reducing the radio interference. When properly installed and serviced they will discharge, without appreciable interference, up to

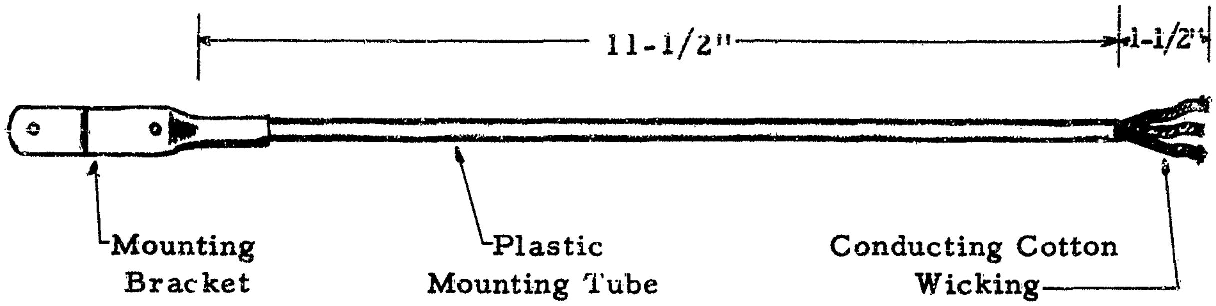


Fig. 4.3.8-A Static Discharger AN/ASA-3

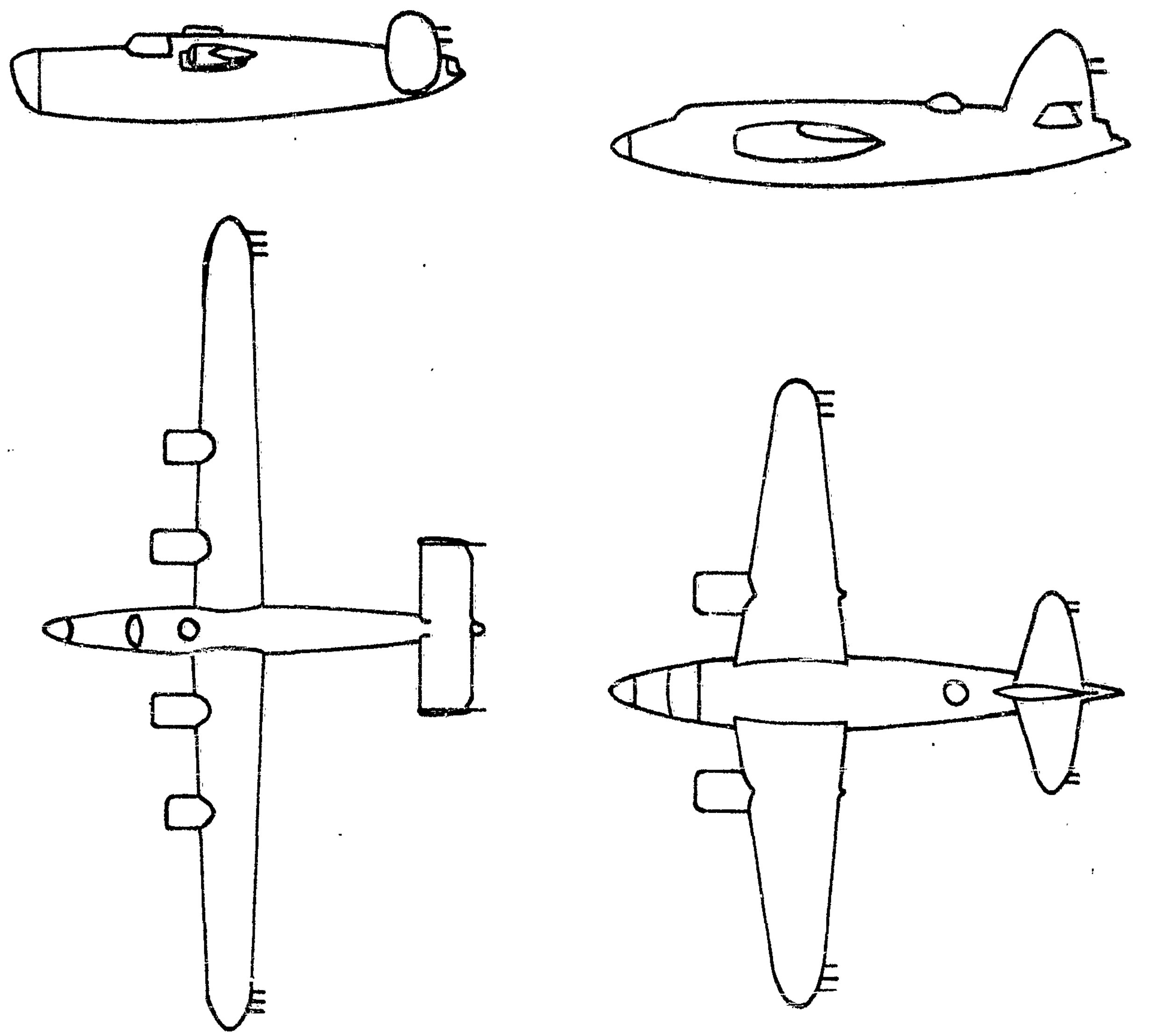


Fig. 4.3.8-B Suggested Locations of Static Discharger AN/ASA-3

250 microamperes per wick.

After exhaustive tests by both civilian and military personnel the "Static Discharger AN/ASA-3" has been officially adopted for use on military aircraft. This discharger consists of a conducting cotton wick 13 inches long and enclosed in a plastic sheath. At one end is an aluminum tube which fits tightly around the wick. The outer end of this tube is flattened to provide for two mounting holes. In operation, the plastic sheath is removed from 1-1/2 inches of the free end of the wick, as shown in Figure 4.3.8-A.

As shown in Figure 4.3.8-B, the dischargers are mounted on the trailing edges of wings, rudders, and stabilizers where the potential produced by precipitation static is highest, but in far enough to escape the extreme turbulence at the wingtips. They should not be placed where oil spraying from an engine exhaust may produce matting of the fibers. At least sixteen dischargers are required per airplane, depending on the size, speed, and type of airplane. The only servicing required is to keep 1-1/2 inches of clean cotton wick exposed outside the plastic sheath at the free end. Replacement should be made when the wicks become shortened by wear to a length of 6 inches.

4.3.9 BLOCK AND SQUIRTER DISCHARGERS

It has been repeatedly observed that precipitation static charges on an airplane can be reduced by use of the radio transmitters in the plane so that radio signals can be received for a short time, even through the most serious interference, immediately after transmitting. It has also been noted that following a lightning discharge in the vicinity of a plane under thunderstorm conditions there are a few seconds of comparative quiet during which signals come through clearly. The explanation is that the additional potential applied to the plane by the transmitter adds a sufficient impulse to discharge the accumulated negative electricity, which is responsible for the precipitation interference. This is because the plane acts as ground for the radio transmitter while its antenna is highly insulated from the remainder of the aircraft. The additional potential given the plane and antenna during the transmission helps to discharge the accumulated static down to a level low enough for the signals to get through. In like manner, the presence of a thunder cloud produces increased negative potentials on one wing and increased positive on the other, both of which may be above the corona point. When the lightning discharge occurs, the space charges within the cloud are reduced and the induced charges on the airplane neutralize each other to such an extent that signals can get through.

Making use of this principle, attempts have been made by several experimenters to produce a system within the plane by which the receiving circuits are momentarily blocked while additional potentials are applied to the plane, or to special dischargers attached to the plane. The time interval during which the receivers are blocked may be three or four times as great as the listening time and is usually adjustable, but there still is a sufficient interval of listening time to obtain range signals clearly. Tests have proved that such a system is expensive to operate. It adds considerable weight and a fire hazard due to the high potentials, and is the equivalent of only two of the AN/ASA-3 dischargers in reducing the interference.

FILTER - A four-terminal network designed to freely transmit currents or voltages of certain frequencies while attenuating all others. - Page 1-49

FREQUENCY TRANSLATION - The production of new frequencies in a non-linear element. - Page 1-3

GROUND - A point of "zero" or "reference" electrical potential, often used in the following sense: (1) To connect to the aircraft structure through a low impedance path, (2) To make equipotential with all other "ground points" in the system. Page 1-45

IMPEDANCE CONCEPT - Consideration of impedance as the ratio of cause to effect leads to the idea which regards the entire aircraft as a single network. - Page 1-19 Appendix V

IMPEDANCE (referring to networks) - The ratio of voltage to current.
IMPEDANCE (referring to media) - The ratio of electric to the magnetic field intensity. - Page V-12

IMPROVEMENT THRESHOLD - The minimum signal-to-interference ratio necessary at the input to produce an intelligible signal at the output. - Page 1-37

INSERTION LOSS - The amount, usually expressed in decibels by which the current in a transmission line, on the load side of the network, has been changed by the insertion of the network. - Page 3-21

INTERFERENCE - VOLTAGE REDUCTION FACTOR - The ratio of the signal-to-interference ratio at the output to that at the input of a receiver. - Page 1-37

INTRINSIC IMPEDANCE - The ratio of electric to magnetic field intensity in a medium in which no reflected wave is present. - Page XVI-2

LABORATORY TESTS - Are measurements of radiated or conducted radio interference in which the test item is placed in a screened laboratory room or in a confined area of low ambient interference under controlled conditions.

MAJOR UNIT - Is an assembly of parts, connected mechanically or electrically, such as a radar transmitter or a power pack, to perform a specific function.

MICROVOLTS PER kc - Interference intensity in microvolts per kc is equal to the number or r.m.s. sine wave microvolts (unmodulated), applied to the input of the measureing circuit at its center frequency, which will result in peak response in the circuit equal to that resulting from the interference pulse being measured, divided by the effective bandwidth of the circuit in kilocycles. The effective bandwidth is the area divided by the height, of the voltage-response-versus-radio-frequency selectivity curve, from antenna to peak detector.

MISMATCH RATIO - The ratio of impedances looking to the right and to the left of a pair of terminals. - Page V-11

NON-LINEAR IMPEDANCES - Impedances that vary with current through them or

voltage across them. - Page 1-12

OPEN SPACE - Is a site ideally in open, flat terrain 100 feet or more from buildings, trees, power lines or communication lines, underground cables and similar obstructions.

PARASITIC OSCILLATIONS - Oscillations which occur at other than a desired frequency or its harmonics, or outside a tank circuit. - Page 3-99

PRECIPITATION STATIC - Radio interference experienced when the flight path is through precipitation. - Page 4-1

RADIATION - The phenomenon of electromagnetic waves spreading out in space from a source according to the laws of wave propagation. - Page 1-28

RADIO INTERFERENCE - Any electrical disturbance which causes an undesirable response or malfunctioning in any electronic equipment. - Page viii Introduction - Page 1-1

RANDOM NOISE - An electrical disturbance that is completely without regularity in its detailed properties. - Page 1-1

RECEIVER - Any electronic equipment in which unwanted signals may cause an undesirable response. - Page viii Introduction.

SHIELD - A partition between two regions of space such that the electric and magnetic fields of interest are attenuated in passing from one region to the other. Page 1-46

SHOT EFFECT - The irregularity of plate current in a vacuum tube due to variations in cathode emission, - Page 1-1

SKIN EFFECT - The crowding of current toward the surface, or skin, of a conductor. - Page 1-48

SPURIOUS RESPONSE - To minimize cross modulation and overloading, good engineering practice requires that one or more tuned circuits shall be placed ahead of the first RF amplifier stage of an interference meter. The meter should also be capable of rejecting spurious responses resulting from combinations with the fundamental or harmonics of the conversion oscillator system of the superheterodyne section of the meter. The degree of rejection is measured in terms of attenuation in db relative to the desired signals. This spurious response rejection should be at least 40 db.

SURFACE CONTACT TRANSIENTS - Transients resulting from the variation in contact resistances across sliding surfaces of rotating electrical machines. Page 3-75

SURFACE TRANSFER IMPEDANCE - The ratio of longitudinal voltage drop along the outside of a tubular shield to the current carried by the shield. - Page XI-2

SYSTEM - Contains two or more Sets or Major Units located at different points but accomplishing their objective through interdependent or interrelated operations, as for example a Propeller Control System.

THERMAL AGITATION - The thermal motion of the conduction electrons in a resistor causing minute interfering currents. - Page 1-1

TO BOND - To connect between two points through a low impedance path. Page 1-40

TRANSMISSION FACTOR (referring to networks) - The ratio of the voltage in the transmitted wave to that in the incident wave at a point of discontinuity.
TRANSMISSION FACTOR (referring to media) - The ratio of the electric field intensity in the transmitted wave to that of the incident wave at a surface of discontinuity. - Page XVI-5

TRANSMIT - RECEIVE (TR) BOX - A device used in radar sets to prevent the transmitted pulse from entering the receiver. - Page 3-105

UNDESIRABLE RESPONSE - Any audible, visible, or otherwise measurable response of a receiver not produced by a desired signal provided that either its duration is longer than one second or its highest recurrence rate during normal operation of the aircraft is greater than once every three minutes. - Page 1-1

WAVE TRAP - A circuit designed to attenuate greatly one frequency or a very narrow band of frequencies while passing without appreciable attenuation all other frequencies. - Page 1-61

APPENDIX III

SELECTED REFERENCES

Paragraph	Title
1	THEORY OF INTERFERING SIGNALS
1.1	STATISTICAL ANALYSIS

(a) Ballantine, S., "Shot Effect in High-Frequency Circuits", JFI, v 206, p. 159-167 (1928).
(b) Bell, D. A., "Shot Noise and Valve Equivalent Circuits", Wireless Eng. v 20, p. 538 (1943).
(c) Carson, J. R., "The Statistical Energy-Frequency Spectrum of Random Disturbances", Bell System Technical Journal, Vol. 10, pp. 374-381, July, 1931.
(d) Glas, E. T., "Frequency Analysis in the Investigation of Radio Interference", Teknisk Tidskrift, Vol. 66 No. 10, Sup. p 41-46, March, 1936.
(e) Goldman, Stanford, "Frequency Analysis, Modulation, and Noise", McGraw-Hill, 1948.
(f) Lawson and Uhlenbeck, "Threshold Signals", MIT-Radiation Laboratory Series, Vol. 24, McGraw-Hill, 1950.
(g) Lee and Stutt, "Statistical Prediction of Noise", TR 129, MIT, July 12, 1949.
(h) MacDonald, D. K. C., "Some Statistical Properties of Random Noise", Proc. Cambridge Philo. Soc., pp. 369 - 372, July, 1949.
(i) Radio News, "Analysis of Radio Interference Phenomena", Vol. 35, p. 54, June, 1946.
(j) Rice, S. I., "Mathematical Analysis of Random Noise", Bell System Technical Journal, 23 - 282 - 232 (July, 1944), 24 - 47 - 156 (Jan., 1945), 621 - 35: 530, 162.

1.2 THE NATURE OF INTERFERENCE

(a) Brooks, Harvey, "Theory of Random and Modulated Noise", NDRC 147.
(b) Franz, K., "Amplitude of Noise Voltages", Wireless Eng., v. 20, p. 397 (1943).
(c) Freberg, C. R., "Noise, Its Character and Measurement", Refrig. Eng., v 53, p. 119 (1947).
(d) Gardner, M. F., Barnes, J. L., "Transients in Linear Systems", John Wiley & Sons, (1942).
(e) Johnson, Walter C., "Transmission Lines and Networks", McGraw-Hill, 1950.
(f) Landon, V. D., "A Study of the Characteristics of Noise", Proc. Inst. Radio Engineers, (1936), 24, 1514-1521.

Paragraph	Title
	(g) Ritz, W. A., "Analysis of Radio Interference Sources", Electron 5:1-13, October, (1949).
	(h) Surdin, M., "Distribution in Time of Spontaneous Fluctuations Voltage", Phil. Mag., 1943, 34 - 716 - 722.
	(i) See Par. 1.1 (e)
	(j) See Par. 1.1 (f)
	(k) See Appendix IV.
1.3	THE ORIGIN OF INTERFERING SIGNALS
	(a) Farrel, O. P., "New Concepts on the Causes of Radio Interference", Radio 29; 27-28 Dec., 1945, Radio 30; 25, Jan., 1946, Radio 30; 23-24; Feb , 1946.
	(b) Shetzline, R. A., "Factors Controlling Man-made Radio Interference", Bell Lab., Rec. 20, pp. 251-253; June 4.
1.3.1.1	ROTATING MACHINERY
	Motter, D. P., "Commutation of DC Machines and Its Influence on Radio-Interference Voltage Generation", Trans. AIEE, p. 491-5, and 495, 1949.
1.3.1.3	NON-LINEAR IMPEDANCE
	(a) Bunimovich, V. I., "Effect of the Fluctuations and Signal Voltages on a Nonlinear System", J. Phy., USSR v 10, n 1. p. 35-48 (1946).
	(b) Everitt, W. L., "Communication Engineering", McGraw-Hill, 1937.
	(c) Middleton, D., "Some General Results in the Theory of Noise Through Non-Linear Devices", Quart. of Appl. Math., 5, 4, 1948, p. 445.
1.3.2.1	BRUSHES
	(a) Hoyes, Pitman, "Current-Collecting Brushes in Electrical Machines", Sir Isaac Pitman, London, 1947.
	(b) Kloeffler, R.G., Kerchner, R.M., Brenneman, J.L., "Direct Current Machinery", The Macmillan Company, 1948.
	(c) Langsdorf, A.L., "Theory of Alternating Current Machinery", McGraw-Hill, 1947.
1.3.2.2	MECHANICAL SWITCHES AND COMMUTATORS
	(a) Foerlag, Gebers, "Electrical Contacts", Ragnar Holm, Stockholm, Sweden, 1946.
	(b) P. R. Mallory & Co., "Mallory Electrical Contacts Data Book", Indianapolis, 1945.
	(c) Olney, F. D., "Resistance Commutation", Trans. AIEE, 1950.

Paragraph	Title
	(d) See Par. 1.3.2.1 (b)
1.3.2.4	ARCING
	Lightning and Transients Research Laboratory, "Radio Interference Transients Investigation II", Filed in Lightning and Transients Res. Lab. Report No. 3.
1.4	CONSIDERATION OF THE ENTIRE AIRCRAFT AS A SINGLE NETWORK
	(a) Navaer, "Elimination of Radio Interference in Aircraft", Report No. 16-5Q-517.
	(b) Shea, T. A., "Transmission Networks and Wave Filters", D. Van Nostrand Co., 1929.
1.5	APPLICABILITY OF THE "IMPEDANCE CONCEPT" TO THE SOLUTION OF INTERFERENCE PROBLEMS
	(a) See Par. 1.4 (b)
	(b) See Par. 1.2 (d)
	(c) See Par. 1.2 (e)
1.6	METHODS OF TRANSMISSION
	Kimbark, E.W., "Electrical Transmission of Power and Signals", John Wiley & Sons, 1949.
1.6.4	RADIATION
	(a) Air Technical Services Command, "Direction of Currents on an Airplane Excited by Radiation", AAF Translation.
	(b) Chatterjee, S.K., "Absorption and Reflection of U.H.F. Waves (300-500 mc/s) by Seawater", Ind. Journal of Phis. June, 1949, (23-6 monthly), pp. 273-279.
	(c) Jordan, E.C., "Electromagnetic Waves and Radiating Systems", Prentice-Hall, 1950.
1.7.1	THE NUISANCE VALUE OF INTERFERING SIGNALS
	Goldman, Stanford, "Frequency Analysis, Modulation, and Noise", McGraw-Hill, 1948.
1.7.2	PULSE LENGTHENING
	See Par. 1.7.1
1.7.4	EFFECTS OF DIFFERENT TYPES OF MODULATION
	(a) OSRD, Studies on the Effect of Noise on Speech Communications", OSRD - 2038, 11/25/43, PB22907.

Paragraph	Title
	(b) OSRD, "Speech in Noise: A Study of the Factors Determining its Intelligibility", OSRD - 4023, 9/1/44, PB19805.
	(c) OSRD, "Effects of Amplitude Distortion Upon the Intelligibility of Speech", OSRD - 4217, 11/15/44, PB19775.
	(d) OSRD, "The Masking of Signals by Noise", OSRD-5387, 10/1/45
	(e) See Par. 1.7.1.
1.8	BASIC METHODS FOR SUPPRESSION OF RADIO INTERFERENCE
	(a) Aggers, C.V., "Methods of Controlling Radio Interference", Electrical Engineering.
	(b) Mallory, P.R. & Co., "Practical Radio Noise Suppression", MYE Technical Manual, p. 309-324, Sec. 10, 1942.
1.8.1	SUPPRESSION AT THE SOURCE
	See Par. 3.2.3 (d)
1 8.1.1	BONDING
	Signal Corps., "Radio Interference Suppression System for 100 kw Engine-Generator Unit", Manufactured by Cummins Engine Co., Inc., Columbus, Indiana, Hall Tech., Memo No. M-1301, SCP No. 4114B - 1085, June 27, 1950.
1.8.1.2	SUPPRESSION OF ARCS
	See Par. 3.2.3 (d)
1.8.2.2	SHIELDING
	(a) Air Technical Services Command, "Shielding Effect of Interference, Eliminating Flexible Tubes", AAF Translation, April, 1946.
	(b) Air Technical Services Command, "Examination of Lines Variously Shielded to Avoid Radio Interference", AAF Translation.
	(c) Breeze Corporation, "Investigation of Shielding Efficiency Theoretical and Practical Investigation of Shielding Efficiency with Separate Measurements of E and A", Terminal Report.
	(d) Levy, S., "Electromagnetic Shielding Effect", Proc. I.R.E., v. 24, #6:923 - 941, June, 1936.
	(e) Naval Research Lab., "Shielding Efficiency of Shielded Enclosures", January 10, 1951.
	(f) Schelkunoff, S.E., "Electromagnetic Waves", D. VanNostrand Co., Inc., 1943, p. 303 - 315.
	(g) See Par. 1.8.1.1.
	(h) See Par. 3.3.2.4 (a)
	(i) See Appendix XVI (a)
	(j) See Appendix XVI (b)

Paragraph	Title
1.8.2.3	FILTERING
	(a) Mason, Warren P., "Electromechanical Transducers and Wave Filters", D. VanNostrand Co., 1946. (b) See Par. 1.8.1.1 (c) See Par. 1.4 (b)
1.8.3	SUPPRESSION IN RECEIVERS
	(a) Pyle, H., "Eliminating Noise in Aircraft Radio Receivers", Aero Eng. Rev. 4; 87, April, 1945. (b) St. Dallas, G., "New Methods for the Reduction of Interference in Receivers", Funktech Monatshefte, No. 5, pp. 69-75, May, 1941.
1.8.3.2	BANDWIDTH CONSIDERATIONS
	(a) Proc. I. R. E., "Relations Between Speed of Indication, Bandwidth & S/N in Radio Navigation and Direction Finding", IRE, May, 1949. (b) See Par. 1.7.1.
2	THE MEASUREMENT OF INTERFERING SIGNALS
	(a) Bond, F. E., "Measurements of Noise, Crosstalk, Volume and Related Quantities in Communication Systems", AMC, USAF, Wright Field, Signal Corps Proj. No. 2051-18, February 24, 1949. (b) ENT, "Measurements of High Frequency Interference", Vol. 18, #6 pp. 126-133, June, 1941. (c) Evans, J., "Measurement and Suppression of Radio Interference", British Institution of Radio Engineers, February, 1949. (d) Fletcher, H., "Noise Measurements", Elec. Engineering, v. 52, p. 744, November, 1933. (e) Franks, C.J., "The Measurement of Radio Noise Interference", RMA Engr., Vol. 3, p. 7-10, November, 1938. (f) Hamburger, G. L., "Interference Measurement", Wireless Engineer, v. 25, n 293 and 294, p. 44-54 and 89-97, February and March, 1948. (g) Kaye, G. W. C., "Measurement of Noise", Phys. Soc. Reports, v. 3, p. 130 (1936). (h) Naday, G., "Noise Measurements in High Frequency Receivers", Annales de Radioelectricite, July, 1949, pp. 257-260. (i) OSRD, "Investigations of the Measurements of Noise", Final report, PB19657, 10/31/45, OSRD-6358. (j) See Par. 1.2 (f). (k) See Par. 3.1.

Paragraph	Title
2.1.1	MILITARY SPECIFICATION MIL-I-6051 MIL-I-6051.
2.1.2	MILITARY SPECIFICATION MIL-I-6181 MIL-I-6181 - Sec. 3.1, 3.2.1 to 3.2.5.
2.1.2.1	SUSCEPTIBILITY LIMITS AND TESTS MIL-I-6181 - Sec. 3.4.1, 3.4.2.
2.1.3	MILITARY SPECIFICATION JAN-I-225 JAN-I-225 - Sec. A, B, C, D, H-1. JAN-I-225 - Sec. F-1; E-2e(3); F-4b(1) to F-4c(4). Also Fig. 2.1.3.
2.1.4	MILITARY SPECIFICATION MIL-I-16910(SHIPS) MIL-I-16910(SHIPS)
2.2	RADIO INTERFERENCE MEASURING SETS (a) Aggers, C.V., Foster, D.E., Young, C.S., "Instruments & Methods of Measuring Radio Noise", AIEE Transactions, Vol. 59, pp. 178-192, March, 1940. (b) Burrill, C.M., "New Equipment to Measure Intensity of Radio Noise", Broadcast News #32, pp. 6-7, March, 1940. (c) Daniel, L.H., "Apparatus for the Measurement of Interference at UHF", ERA Report M/T 64, 1939. Also, IEE Journal, Vol. 88, Part III, pp. 41-49, March, 1941. (d) Davis, A.H., "An Objective Noise Meter for the Measurement of Moderate and Loud, Steady and Impulsive Noises", Journal IEE, Vol. 83, pp. 249-260, August, 1938. (e) OSRD, "Standard Radio Noise Signal Generator", OSRD-5284.
2.2.2.1	RADIO TEST SET AN/URM-6 General Electric, "Measurement of Radio Noise with a Radio Receiver", May 21, 1948.
2.2.2.5	RADIO TEST SET AN/PRM-1 (a) NAVSHIP-91255 - "Instruction Book for Radio Test Set, AN/PRM-1", Navy Dept. Publication, dtd. 30 September 1949. (b) See Par. 2.2.4.1.
2.2.2.6	RADIO INTERFERENCE TEST SET AN/URM-3 (a) Coles Signal Lab., "Test Set AN/URM-3( ), under Development by Coles Sig. Lab., 5/20/46.

Paragraph	Title
	(b) See Par. 2.2.4.1.
2.2.2.7	MEASUREMENTS MODEL 58
	Fowler, Nicholson, Krell, "Investigations of the Measurements of Noise", Progress Report No. 15, Univ. of Penn. Research Division. January 15, 1950, Report #50-15.
2.2.2.8	NOISE-FIELD INTENSITY METER TS-587/U and TS-587A/U
	University of Pennsylvania, "Investigations of the Measurement of Noise", Progress Report No. 11, November 15, 1948, SER #49-7.
2.2.2.11	RADIO-INTERFERENCE MEASURING SET AN/URM-29
	See Par. 2.2.4.1.
2.2.4	SELECTION OF TEST LOCATIONS
	MIL-I-6051 - Sec. 4.2.3.4 to 4.2.3.6.
3	APPLICATION OF THEORY TO PRACTICE
	(a) See Par. 2(b). (b) See Par. 2(c).
3.1	GENERAL DESIGN CONSIDERATIONS FOR INTERFERENCE-FREE OPERATION
	War Dept., "Elimination of Radio Noise in Aircraft", (I) Int. to aircraft radio noise (II) Detecting, localizing and measuring radio noise; (III) Application of suppression techniques, June 25, 1945.
3.1.1	FILTERS AND OTHER SUPPRESSION NETWORKS
	(a) Corby, R.E., "The Theory of a Three-Terminal Capacitor", Proc. IRE, June, 1950, Vol. 38, No. 6, p. 635-6. (b) Frick & Zimmerman, "Aircraft Radio Noise Filters", Aircraft Communications 23, 32, 42, July, 1943. (c) Rollin Company, "Development of High Efficiency Low Pressure Radio Filters", Lab. Report, Contract W33-038-AC-17991, December 15, 1947.
3.1.1.2	POWER LINE FILTERS
	(a) AEEL, "Design Details of a 500 amp. 50 volt Line Filter Suitable for Use in Naval Aircraft", Report No. 75-48, September 7, 1948. (b) Edwards, Jonathan, "An Electronic Power Supply Filter", IRE

Paragraph	Title

Fall Convention, San Francisco, 1949.
(c) See Par. 3.1.3.

3.1.1.3 HARMONIC SUPPRESSION FILTERS

(a) Linder, L., "The Effect of Self-Inductance of Rolled Paper Capacitors on Their Impedance", Elektrotech, Zeit, Vol. 60, pp. 793-798, AF Translation dtd. August, 1946.
(b) Mason, M.P., "Filters & Transformers Using Coaxial & Balanced Transmission Lines", Bell System Technical Journal, Monograph B-1003.
(c) See Par. 3.1.1.

3.1.1.5 FEED-THROUGH CAPACITORS

Sprague Electric Co., "Small Size Feed-Through Capacitors", Oct. 18, 1948.

3.1.2 SHIELDING

(a) AMC, "Radio Shielding of Aircraft Engine Ignition Systems", A.A.F., Engineering Division, Technical Note TN-92-1.
(b) Anderson, Alton R., "Cylindrical Shielding and Its Measurement at Radio Frequencies", Proceed. of the IRE and Waves and Electronics, Vol. 34, No. 5, May, 1946.
(c) ATSC, "Shielding Effect of Interference Eliminating Flexible Tubes", A.A.F. Translation No. 518, April 22, 1946.
(d) Lightning & Transients Lab., "Aircraft Interference Fields", Report No. 174, January, 1951, p. 11-12-17-23-24.
(e) Roston, B., "Shielding at VHF", Wireless Engineer, July, 1948, Vol. 25, No. 298, p. 221-230.
(f) Schaefer, "The Effect of Different Types of Electromagnetic Shields", Translation by Univ. of Cincinnati, Edited by Leibiger Research Lab., Inc., August 27, 1947.
(g) Technicraft Lab., "Study of Tubular Shielding", Final Report Vol. I, June 28, 1948; Vol. II, June 29, 1949.
(h) Zepler, E.E., "The Technique of Radio Design", John Wiley and Sons, Inc., Chap. 8, The Principles of Screening, 1943.
(i) See Par. 1.8.1.1.
(j) See Par. 1.8.2.2 (c).
(k) See Par. 1.8.2.2 (f).

3.1.2.3 JOINTS

See Par. 3.1.2(h).

3.1.2.6 CONTACT IMPEDANCE

(a) Curtis, A.M., "Contact Phenomena in Telephone Switching

Paragraph	Title
	Circuits", Bell System Technical Journal, Vol. 19, No. 1, p. 40-62, Jan., 1940.
	(b) Hunt, L. B., "Electrical Contacts", Johnson, Mathey & Co., Ltd., & Mallory Metallurgical Prod., Ltd., London, 1946.
3.1.2.7	CORROSION
	Germer, L. M., Haworth, F. E., "Erosion of Electrical Contacts on Make", Journal of Applied Physics, Vol. 20, p. 1058-1109, November, Bell Telephone System Monograph No. 1710.
3.1.2.10	RIGID CONDUIT
	(a) Browning, G.H., "Reducing Man-Made Static (The Use of a Shielded Lead-In and the Calculation of the Appropriate Impedance-Matching Network)", W.E. & E.W. 10, 157; March, 1933, Electronics, pp. 366-368, December, 1932.
	(b) Shive, S. L., "Effectiveness of Conduit as R-F Shielding", Electronics, February, 1946, 160-166. (Ref.: Coles Signal Lab. Tech. Memo No. M-1019 of 10-21-46, Dev. of Procedure for Testing R-F Leakage in Shielding Conduit; and U. S. Patent 2, 446, 195 of 8-3-48, Tester for Electrical Shieldings.)
	(c) See Par. 3.1.2 (f).
	(d) See Par. 3.1.2 (g).
3.1.2.11	FLEXIBLE CONDUIT
	(a) ATSC, "Shielding Effect of Interference Eliminating Flexible Tubes", PB23544, AAF Translation.
	(b) See Par. 3.1.2 (f).
	(c) See Par. 3.1.2 (g).
	(d) See Par. 3.1.2.10 (a).
	(e) See Par. 3.1.2.10 (b).
3.1.3	BONDING
	(a) Micro Circuits Co., "Non-Hardening Silver Micropastes", January, 1951.
	(b) Signal Corps, Technical Report No. T-1272, Fort Monmouth, N. J., November 19, 1951.
	(c) Watson, Randolph, "Aircraft Electrical Engineering", McGraw-Hill Book Co., 1943.
	(d) See Par. 1.4 (a).
	(e) See Par. 1.8.1.1.
	(f) See Par. 3.1.2.6 (b).
3.1.3.1	DIRECT BONDING
	Blake, K.W., "Corrosion and Radio Interference", Royal NavSci.

Paragraph	Title
	Ser J. 1, 34-36, September, 1945.
3.1.3.2	BONDING JUMPERS
	Whitaker Cable Corp., "Qualification Test of Bonding, Quick Disconnect, and Current Return Jumpers Types I, II, and III", ENRE, USN, ADS, October 18, 1949.
3.1.4	SPECIAL CIRCUITS
	(a) Carson, J.R., "Selective Circuits and Static Interference", Trans. AIEE v 43, p. 789, 1924. (b) Lightning & Transients Research Institute, Technical Report No. 7-A, Minneapolis, 1948.
3.1.4.1	LIMITERS
	(a) Haviland, R.P., "Simple Noise Limiter", Radio & Television News 44, August, 1950. (b) Maguire Industries, "Noise Limiting Circuits". (c) Markus, John, and Zeluff, Vin, "Handbook of Industrial Electronic Circuits", Chapter on Limiter Circuits, p. 85, McGraw-Hill, 1948. (d) Terman, F. E., "Radio Engineers Handbook, Sec. 9., Par. 6-8, McGraw-Hill, 1943. (e) U. Radio, "Noise Limiters (Manual & Automatic Types for UHF and BC Receivers)", #268, p. 26-28, May, 1942. (f) See Par. 1.4 (a), Sec. X, Par. 2.
3.1.4.2	WAVE TRAPS
	See Par. 1.4 (a), Sec. XIV, Par. 8.
3.2	DESIGN CONSIDERATIONS APPLIED TO AIRCRAFT COMPONENTS FOR MINIMUM GENERATION OF INTERFERENCE
	(a) Frick, C., "How to Minimize the Radio Influence of Electric Devices", Electrical Mfg., March, 1939. (b) Gill, A.J., Whitehead, S., "Electrical Interference with Radio Reception", JIEE, p. 345-386, 1938, Vol. 83. (c) Kogel, K., "Methods of H. F. Interference Suppression for Machines & Appliances up to 500 Watts", AEG Mitteilunger, pp. W238, September/October, 1941. (d) Majumdar & Khastgir, "Studies of Electrical Interference to Radio Reception", Ind. J. Phys., Vol. 17, p. 271-282, October, 1943.
3.2.1.1	ROTATING MACHINERY
	(a) Frick & Motter, "Investigation of Radio Noise Generation in

Paragraph	Title
	Aircraft Electrical Machinery", G.E. Terminal Report, Contract W28-099-ac-83.
	(b) Signal Corps, "Suppression of Radio Interference Created by Engine-Generator Units", Manual SIG-461-1, August, 1945.
3.2.1.1.1	BRUSHES
	(a) See Par. 3.2.1.1 (a).
	(b) See Par. 1.3.2.1 (a).
3.2.1.1.2	COMMUTATION
	See Par. 3.2.1.1 (a).
3.2.1.1.4	SHIELDING
	(a) Diamond, H., Gardner, F.G., "Engine-Ignition Shielding for Radio Reception in Aircraft", Proc. I.R.E., v. 18 #5, 840-861, May, 1930.
	(b) See Par. 3.3.2.4 (a).
3.2.1.1.8	ALTERNATORS
	Naval Air Material Center, Phila., Pa., "Radio Interference Correctives for 800-1 Series Motor Alternators", Project TED NAM EL-406, Report No. ARRL 511-47.
3.2.1.2	VIBRATORS
	ASTM, "Bibliography and Abstracts on Electrical Contacts", 1944.
3.2.2	ELECTRICAL OSCILLATIONS IN STATIONARY CIRCUITS
	(a) Carson, J.R., Zobel, O.J., "Transient Oscillations in Wave Filters", Bell System Technical Journal, July, 1923.
	(b) Eastman, A.V., "Fundamentals of Vacuum Tubes", McGraw-Hill, Chap. XI, 1941.
	(c) Glasgow, R.S., "Principles of Radio Engineering", McGraw-Hill, Chap. V., 1936.
3.2.2.1	LOCAL OSCILLATORS
	See Par. 3.3.2.
3.2.2.2	TRANSMITTERS
	(a) Black, H.S., "Stabilized Feedback Amplifiers", Bell System Technical Journal, 13, p. 1, January, 1934.
	(b) Cunningham, W.J., Gofford, S.J., and Licklider, J.C.R.,

Paragraph	Title
	"The Influence of Amplitude Limiting and Frequency Selectivity Upon the Performance of Radio Receivers in Noise", Proc. IRE, v. 35, p. 1021, October, 1947.
	(c) Freeman, R.L., "Use of Feedback to Compensate for Vacuum-Tube Input-Capacitance with Grid Bias", Proc. IRE, Vol. 26, pp. 1360-1366, November, 1936.
	(d) Terman, F.E., "Radio Engineering", McGraw-Hill, Chap. VI., 1947.
	(e) See Par. 3.2.2 (b), Chap. IX.
	(f) See Par. 3.4.3.
3.2.2.3	MODULATORS
	(a) Claview, A. G., "Signal-To-Noise Radio Improvement in a Pulse-Count-Modulation System", FTL, Elec. Communication, pp. 257-262, September, 1949.
	(b) DeLoraine, E.M., "Pulse Modulation", Electrical Communication 26, pp. 222-227, September, 1949.
	(c) Mallach, L. W., "Reduction of Radar-Radio Interference from Modulators", Radiation Labs., MIT, August 3, 1943.
3.2.2.4	TRANSFORMERS
	(a) Legg, V.E., "Magnetic Material and Method of Treatment Thereof", U. S. Patent #1,998,840.
	(b) See Par. 3.1.4.1, Sec. 2, Par. 24.
	(c) See Par. 3.1.2 (h), Chap. 11.
3.2.2.5	PARASITIC OSCILLATIONS
	(a) Hultbert, C.A., "Neutralization of Screen Grid Tubes to Improve the Stability of Intermediate - Frequency Amplifiers", Proc. IRE, Vol. 31, pp. 663-665, December, 1943.
	(b) See Par. 3.1.4.1, Sec. 8, Par. 6.
	(c) See Par. 3.2.2 (f), Chap. XI.
3.2.2.6	TRANSIENTS IN RESONANT CIRCUITS
	(a) Cruft Elec. Staff, Harvard University, "Electronic Circuits and Tubes", Chap. VII, Sec. 4, McGraw-Hill, 1947.
	(b) Kallman, H.E., Spencer, R.E., Singer, C.P., "Transient Response", Proc. IRE, Vol. 33, pp. 169-195, March, 1945.
3.2.3	ARCING
	(a) Hernz, Fisher, and Peger, "Time Function of the Radiation of Short Spark Discharges", ATI, Vol. 13, #13.
	(b) Kleis, J.D., "How Contact Arcing Can Be Minimized", Electrical Manufacturing, January, 1945.

Paragraph	Title

(c) Matson, Randolph, Aircraft Electrical Engineering, p. 222, McGraw-Hill, 1943.
(d) War Dept., "Suppression of Radio Noises", TM 11-483, Nov., 1944.
(e) See Par. 3.2.2, p. 110-111.

3.2.3.1 IGNITION SPARKS

(a) AAF, "Radio Shielding of Aircraft Engine Ignition Systems", Army Air Force Material Command, Serial TN-92-1, p. Xi-Xii, May, 1944.
(b) Bendix Scintilla, "Development of Lightweight Magnetos", 3rd Quarterly Report, SCL-3071, March 15, 1948.
(c) Curtis, L. F., "Ignition Disturbances", Proceedings of the Radio Club of America, October, 1936.
(d) Nethercot, W., "The Relation Between the Sparking Plug Current and the Short Wave Radiation Produced by Ignition Systems", ERA Report M/T 53, 1937.
(e) Pender, Harold, DelMar, Wm.A., Electrical Engineers Handbook, Sec. 17, p. 137, 138, John Wiley & Sons.
(f) Purdy, R.M., "Development of Lightweight Magnetos", Second Quarterly Report.
(g) Sprague Electric Co., "Radio Interference Suppressors", Contract AF 33(038)0353, February-April, 1950.
(h) War Dept., "Radio Interference Suppression System for International Harvester Co. Model UD-16 Engine (Magneto Ignition)", TM No. M-1279.

3.2.3.3 RELAYS

(a) Dallas, J. P., McCully, T. M., "Sensitive Relay Contact Protection", Trans. AIEE, p. 1204-7, 1948.
(b) Lightning & Transients Research Lab., "Radio Interference Transients Investigation", Technical Report No. 2, p. 1-32, March 15, 1946.
(c) Lightning & Transients Research Lab., "Radio Interference Transients in Aircraft - I", Technical Report, 1949.
(d) Packard, Charles A., "Relay Engineering", Struthers-Dunn, Inc., Philadelphia, 1945.

3.2.3.4 POOR BONDING

(a) See Par. 1.3.2.2 (a).
(b) See Par. 1.3.2.2 (b).

3.2.3.5 SWITCHES

Pfann, W. G., "Bridge Erosion in Electrical Contacts and Its Prevention", Trans. AIEE, p. 1528-33, Bell Telephone System Monograph B-1621.

Paragraph	Title
3.2.3.6	FLUORESCENT LAMPS

(a) Young, C.S., "Radio Noise from Fluorescent Lamps", EEI Bull. Vol. 12, p. 59, February 2, 1944.
(b) See Par. 3.2.3 (e), p. 15-29.

3.3 DESIGN CONSIDERATIONS APPLIED TO AIRCRAFT SYSTEMS

Busignies, H., "Some Relations Between Speed of Indication Bandwidth and Signal-to-Random Noise Ratio in Radio Navigation and Direction Finding", Proc. IRE, Vol. 37, p. 478-488, May, 1949; Electrical Communication, Vol. 26, p. 228-242, September, 1949.

3.3.1.6 LORAN SYSTEM

(a) AMC, "Interference Rejection Antenna System for Loran AN/APN-4 and Equipments", Memo TSELR-66.
(b) Pierce, McKenzie, and Woodward, "Loran", MIT Series, Vol. IV.

3.3.1.8 NAVIGATIONAL RADAR SET

Roberts, A., "Radar Beacons", Chap. 4, Vol. 3, Rad. Lab. Series, McGraw-Hill, 1947.

3.3.1.12 GUN-LAYING RADAR SYSTEM

(a) ATSC, "Notes on Design of Radar Systems which do not Cause Radio Interference", Tech. Note TN-TSERP2-1, February 20, 1945.
(b) Cady, Karelitz and Turner, "Radar Scanners and Radomes", MIT Series, Vol. 26.
(c) MIT, "Interference Between SCR-584's Tracking APN-19 Beacon", MIT 816.

3.3.2 ELECTRICAL SYSTEMS

O'Neil, "Radio Interference Suppression - Investigation of 24-volt Waterproof Electrical System", Tech. Memo No. M-1261, SCP No. 4114A-1145, March 10, 1950.

3.3.2.1 AC AND DC POWER SYSTEM

Weinstein, G., Howell, H.H., Lowe, G.P., Winter, B.J., "Radio-Noise Elimination in Military Aircraft", Trans. AIEE, p. 793-5, 1944.

3.3.2.4 SPARK-TYPE IGNITION SYSTEM

(a) AAF, "Radio Shielding of Aircraft Engine Ignition Systems",

Paragraph	Title
	TN-92-1, May 26, 1944 and July 1, 1946.
	(b) AMC, "Ignition System Exciter", September 29, 1944, 92-885.
	(c) Inst. of Aero Sc., "Radio Noise in Aircraft Engines", Rev. 4, Aero Eng. 13-32, August, 1945.
	(d) Nethercot, W., "Car Ignition Interference", Wireless Engineer, Vol. 26, No. 311, p. 251-255, August, 1949.
	(e) Pierce, R.H., "Radio Interference Suppression System for Cushman Models M6, M7-2, M7-3, and M8 Gasoline Engines", Signal Corps Labs., SCP 4114C-2171, May 8, 1950.
	(f) Wolcott, L.C., "Airplane Motor Radio Shielding", Aviation Engineering, June, 1931.
	(g) See Par. 3.2.1.1.4.
3.4	DESIGN CONSIDERATIONS FOR AIRCRAFT RECEIVERS
	(a) Goldman, Stanford, "Some Fundamental Considerations Concerning Noise Reduction and Range in Radar and Communication", Tech. Report No. 32, Res. Lab. of Electronics, MIT, December 15, 1947.
	(b) OSRD, "Auditory Tests of the Ability to Hear Speech in Noise", OSRD-3516, September 1, 1944.
3.4.1	PATHS OF ENTRY
	(a) Curtiss-Wright Corp., "Investigation of Radio Interference Analysis in a Type B-17 Aircraft", Report TE-49-12-E, Part 3, July, 1950.
	(b) General Lab. Associates, "Factors Influencing Generation of and Susceptibility to Interference Voltages by Electronic Equipment", 1949.
	(c) OSRD, "A.F. Factors in Portable Radio Equipment - Redesign of SCR-274N and ATA Modulators - Performance of Eaves Sound Projector", OSRD-1528, June 28, 1943.
	(d) Royal Aircraft Establishment, "Mutual Interference Between Radio Communications and Radar Equipments", Section 4.
	(e) See Par. 1.4 (a), Sec. X.
3.4.1.1	ANTENNA AND LEAD-IN
	(a) ATSC, "Coupling of Airplane Antennas with the Aircraft Wiring System & the Effect of the Aircraft Wiring System", AAF Trans. PB-25227.
	(b) Burgess, R.E., "Noise in Receiving Aerial Systems", Proc. Physical Society, v. 53, p. 293-304, 1941.
	(c) See Par. 3.4.3.
3.4.1.2	POWER LEADS
	(a) Post Office Engineering Dept., London, England, Radio Report

Paragraph	Title
	No. 2009-1950. (b) See Par. 1.4 (a), Sec. IX.
3.4.1.3	CONTROL CABLES
	(a) See Par. 1.4 (a), Sec. IX. (b) See Par. 3.4.1 (a).
3.4.1.5	CASE SHIELDING
	(a) See Par. 1.4 (a), Sec. IV and X. (b) See Par. 3.4.1 (a). (c) See Par. 3.4.1 (c).
3.4.2	INTERNAL COUPLING
	See Par. 3.4.1 (a), Sec. IV and X.
3.4.3	SPECIAL CIRCUITS
	Naval Research Laboratory, "Spurious Radiation from Radio Transmitters and Receivers via Antenna Paths", BuAer Problem A276R-U, Electronics, NRL Letter 3920 168/48 October 11, 1948.
3.4.3.1	RADIO RECEIVERS
	(a) Admiralty Signal Establishment, "Interference-Free Transmitters and Interference Immune Receivers", Section B, July, 1945. (b) Burgess, R.E., "Receiver Input Circuits, Design Considerations for Optimum Signal/Noise Ratio", Wireless Engineer, v 20, p. 66, 1943. (c) Friis, "Noise Figures of Radio Receivers", Proc. IRE, p. 419 July, 1944. (d) Fubini and Johnson, "Signal to Noise Ratio in A.M. Receivers", Proc. IRE, p. 1461, December, 1948. (e) Goldman, Stanford, "Some Final Consideration Concerning Noise Reduction and Range in Radar and Communication", Proc. IRE, p. 554-594, May, 1948. (f) Millman, Jacob, Seeley, Samuel, Chapter XIX, Sec. 19-10, Electronics, McGraw-Hill. (g) Reich, H.J., "Interference Suppression in AM and FM, Communications, vol. 22, No. 8, p. 7, 16, 19-20, August, 1942. (h) Rogers, D.C., "Suppressing Impulse Noise, Wireless World 55: 489-492, December, 1949. (i) Zinn, M.K., "Transient Response of an F-M Receiver", Bell System Technical Journal, Vol. 27, p. 714-731, October, 1948. (j) See Par. 1.4 (a), Sec. IV and X. (k) See Par. 3.2.2.6 (a), Chapter 23, Sec. 17-22.

Paragraph	Title
	(l) See Par. 3.2.3.1 (b).
3.4.3.2	RADAR RECEIVERS

(a) ATSC, "Notes on Design of Radar Systems Which do not Cause Radio Interference - April 25, 1945.
(b) Beard, E.G., "Reduction of Noise at U.H.F. in Triode Valves and Grounded Grid Amplifiers", Philips Tech. Commun., p. 20-3, March, 1947.
(c) Bell System Technical Journal, "Radar Systems and Components", p. 364-740 - Van Nostrand - 1947.
(d) Goldman, Stanford, "Reduction Radar Noise", Proc. IRE 36, p. 584-594, 1950.
(e) Levy, M., "Signal/Noise Ratio in Radar", Wireless Engineer, Vol. 24, n. 291, p. 349-52, December, 1947, Wireless Engineer Vol. 25, n. 298, p. 236-7, July, 1948.
(f) MIT, "Principles of Radar", Radar School Staff, 2nd Edition, McGraw-Hill Book Co.
(g) Pollard, Ernest, Sturtevant, Julian, "Microwaves and Radar Electronics", Chapter 5 - Circuit Bandwidth, Chapter 8 - I-F Amplifiers, Receiving Gating, Anti-Jamming, and Anti-Clutter Circuits, Chapter 10 - Receiver Gating, Wiley - 1948.
(h) Ridenour, "Radar Systems Engineering", MIT Series, Vol. I.
(i) Royal Aircraft Establishment, "Interference in Communication Receivers from Radar D.C. Pulses", I - Outline of Fourier Analysis, II - Application of Theory of Part I, III - Relationship between Spectral Density Curves and the Interference Curves, IV - Experiment which Verifies the Experimental Result, December, 1945.
(j) Wireless Engineer, 'Signal-Noise Ratio in Radar", p. 399-352, December, 1947.
(k) See Par. 1.4 (a), Sec. X.
(l) See Par. 3.2.2, Sec. 99.
(m) See Par. 3.4.3.1 (e).

3.4.4 DESIGN CONSIDERATIONS FOR MINIMUM OSCILLATOR RADIATION

McClennon, D., "Spurious Radiation from Radio Transmitters and Receivers via Antenna Paths", Final Report, Naval Research Lab.

4.1 CHARACTERISTICS OF PRECIPITATION STATIC

(a) Appleton, E.V., Watson-Watt, R.A., and Herd, J.F., "On the Nature of Atmospherics, II and III", Proc. Royal Soc., 1926, p. 615-77.
(b) Appleton, E.V., and Chapman, F.W., "On the Nature of Atmospherics, IV", Proc. Royal Soc., v. 158, p. 1-22, 1937.
(c) Gunn, R., Parker, J.P., "The High-Voltage Characteristics

Paragraph Title

of Aircraft in Flight", Proc. IRE, Vol. 34, No. 5, p. 241-247, May, 1946.
(d) Lutkin, F.E., "The Nature of Atmospherics VI", Proc. Royal Soc., Vol. 171, p. 285-313, 1939.
(e) Morgan, Proc. I.R.E., p. 969, 1936.
(f) Munro, G.H., and Webster, H.C., "Nature of Atmospherics", Proc. Royal Soc., v. 134, p. 850 and 987, 1934.
(g) OSRD, "Precipitation Static Research", OSRD-1807, August 31, 1943.

4.2 ATMOSPHERIC CONDITIONS FAVORABLE TO PRECIPITATION STATIC

(a) ATSC, "Radio Interferences by Electrical Charges and Their Removal", AAF Trans., PB27606.
(b) OSRD, "Precipitation Static Exploration C-70", OSRD-3595, February 15, 1944.
(c) Malan, D.J. and Schonland, B., "The Distribution of Electricity in Thunderclouds", Proc. Royal Soc. A, 158-177, 1951.
(d) Schonland, B., "The Flight of Thunderbolts", Clarendon Press, Oxford, 1950.
(e) Simpson, G.C., "The Electricity of Atmospheric Precipitation", Phil. Mag. v. 30, p. 1-12, 1951.
(f) Stimmel, R.G., Rogers, E.H., Waterfall, F.E., and Gunn, R., "Electrification of Aircraft Flying in Precipitation Areas", Proc. IRE, Vol. 34, No. 4, p. 167P-177P, April, 1946.
(g) See Par. 4.1 (a).
(h) See Par. 4.1 (b).
(i) See Par. 4.1 (e).
(j) See Par. 4.1 (f).

4.2.2 ICE CRYSTALS, HIGH ALTITUDES, LOW TEMPERATURES

(a) Air Forces Manual No. 40, p. 6, December, 1944.
(b) George, R.H., Tech. Report C.A.A. No. 29, April, 1940, p. 2.
(c) Gunn, R., Proc. I.R.E., p. 177, April, 1946.

4.2.3 FRICTIONAL CHARGING BY IMPACT WITH WATER DROPS

(a) Gott, J.P., "Charge on Water Drops Falling Through a Cloud of Charged Particles in a Vertical Field", Proc. Royal Soc., Vol. 151, p. 665-684, 1935.
(b) Nukiyama, D., "On the Accumulation of Electric Charges on Thunderclouds", Japanese Jour. Astron. and Geophysics, Vol. VI, No. 1, 1928.
(c) Nukiyama, D. and Noto, H., "On the Electrification of Water Drops", Japanese Jour. Astron. and Geophysics, Vol. VI, No. 1, 1928.

Paragraph	Title
	(d) Zeleny, John, "Current-Voltage Characteristics of High-Potential, Direct Current Brush and Glow Discharges in Air at Atmospheric Pressure", Jour. Franklin Institute, p. 685-699, 1934.
4.2.3.3	FLIGHT THROUGH RAIN DROPS
	(a) Humphreys, W.J., Lippincott, "Physics of the Air", p. 314.
	(b) Morgan, H.K., "Rain Static", Proc. I.R.E., Vol. 25, p. 959-963, 1936.
4.2.4	CHARGES INDUCED BY EXTERNAL FIELDS: THUNDERSTORMS
	(a) Banerji, S. K., "The Electric Field of Overhead Thunderclouds", Royal Meteorological Society, J, Vol. 56, p. 305, 1930.
	(b) Baumler, M., "The Simultaneous Appearance of Atmospheric Disturbances", Jb. Drahtl. Telegr., Vol. 22, p. 2-8, 1923.
	(c) Bellaschi, P. L., "Lightning Strokes in Field and Laboratory" Trans. A.I.E.E., August, 1941.
	(d) Chapman, F.W., "Atmospheric Disturbances Due to Thundercloud Discharges", Part I, Proc. Phys. Soc. London, Vol. 51, p. 876-894, 1939.
	(e) deBallore, R., "The Relation Between Storms and Atmospherics", Onde Elect., Vol. 8, p. 463-464, 1929.
	(f) Gould, K. E., "Fields Caused by Remote Thunderstorms", Elect. Eng., Vol. 55, p. 575-582, 1937.
	(g) Halliday, E. C., "The Polarity of Thunderclouds", Proc. Royal Soc. A, p. 205-228, 1932.
	(h) Jensen, J. C., "The Branching of Lightning and the Polarity of Thunderclouds", J. Franklin Institute, Vol. 216, p. 707-748, 1933.
	(i) Morganrath, O., "In Atmospheric Storms", Funktech M., Vol. 3, p. 70-74, 1939.
	(j) Simpson, G. C., "The Mechanism of a Thunderstorm", Proc. Royal Soc., Vol. 114, pp. 376-401, 1927.
4.2.4.2	EFFECTS OF TURBULENCE IN THUNDERSTORMS
	(a) Gen. Elec. Rev., p. 414-418, September, 1936.
	(b) "Terres. Magnetism and Atmospheric Electricity", p. 93, March, 1935.
	(c) Byers, Misc. Report No. 3, Univ. of Chicago, Institute of Meteor.
4.2.4.3	MAGNITUDE OF ELECTROSTATIC FIELDS AND THEIR EFFECTS IN THUNDERSTORM AREAS
	(a) Malan, D. and Schonland, B., "The Electrical Processes in the Intervals Between the Strokes of a Lightning Discharge",

Paragraph	Title

Proc. Royal Soc. A, p. 145-163, 1951.
(b) McEachron, K.B., and Patrick, K.G., "Playing with Lightning", Random House, New York, 1940.
(c) Norinder, H., "Rapid Variations in the Magnetic Field Produced by Lightning Discharges", Proc. Phys. Soc., London, Vol. 49, p. 364-375, 1937.
(d) Schonland, B., "The Polarity of Thunderclouds", Proc. Royal Soc., Vol. 118, p. 233-251.
(e) Simpson, G., and Robinson, O., "The Distribution of Electricity in Thunderclouds II", Proc. Royal Soc., Vol. 177, p. 281-329, 1941.
(f) Wilson, C. T. R., "Investigation of Lightning Discharge and on the Electric Field of Thunderstorms", Philos Trans. (A), Vol. 221, p. 73-115, 1921.

4.2.4.4 LIGHTNING DISCHARGES TO AIRCRAFT

(a) Allbright, J.G., "A Lightning Flash and Its Component Strokes", Journal Applied Physics, Vol. 8, p. 313-318, 1937.
(b) Allibone, T. E., "The Mechanism of the Long Spark", Jour. Inst. Elect. Engineers, London, p. 513-521, May, 1938.
(c) Bruce, C.E.R., and Golde, R.H., "The Lightning Discharge", J. Inst. Elec. Engrs., Vol. 88 (2), p. 487-505, 1941.
(d) Evans, E.A., and McEachron, K.B., "The Thunderstorm", Gen. Elect. Rev., p. 413-425, 1936.
(e) George, R.H., C.A.A. Technical Report No. 9, p.2, April, 1940.
(f) Goodlet, B.L., "Lightning", Journ. Inst. Elec. Engrs., Vol. 81, p. 1-26, 1937. Completed form, Vol. 82, p. 1-55, 1938.
(g) Meek, J. M., "The Mechanism of the Lightning Discharge", Vol. 55, p. 972-977, 1939.
(h) Simpson, G.C., "Lightning", Jour. Inst. Elec. Engrs., Vol. 67, p. 1269-1283, 1929.
(i) See Par. 1.3.2.4.
(j) See Par. 4.2.4.3 (f).

4.3 METHODS FOR REDUCING INTERFERENCE FROM PRECIPITATION STATIC

(a) AN, "Aircraft Treatment for the Elimination of Precipitation Static", AN-P Stat. Proj., Report A955A.
(b) OSRD, "Precipitation Static Reduction C-21", OSRD - 1907, 3/31/43.
(c) Kinser, G.D., "Investigations of Methods for Reducing Precipitation Static Radio Interference", Proc. IRE, Vol. 34, No. 5, p. 234-240, May, 1946.
(d) See Par. 4.2 (a).

Paragraph	Title
4.3.3	DESIGN OF AIRCRAFT STRUCTURES
	(a) Military Specification, MIL-B-5087. (b) See Par. 3.1.4, p. 22 ff.
4.3.4	IMPROVED ANTENNA WIRE AND INSULATORS
	Lightning and Transients Institute, Quarterly Progress Report No. 21693-16, June 16, 1950, p. 5 ff.
4.3.5	LOOP TYPE ANTENNAS
	(a) Chapman, L.A., "Anti-Interference Aerials", Wireless World 38 Vol. 15, p. 3, January, 1936. (b) Crossley, A., "Interference Reducing Antenna Systems", QST No. 26, p. 25-26, May, 1942. (c) Sandretto, "Principles of Aeronautical Radio Engineering", McGraw-Hill, p. 114 ff. (d) Starr, E.C., "Precipitation - Static Radio Interference Phenomena on Aircraft", Bull. 10, 1939, Engineering Experiment Station, Oregon State College.
4.3.6	TRAILING WIRE DISCHARGERS
	(a) National Advisory Committee for Aeronautics, Technical Note No. 1001, 1946. (b) See Par. 4.3.5 (a). (c) See Par. 4.3.5 (b). (d) See Par. 4.3.5 (c).
4.3.7	FLUSH ANTENNAS, CANOPIES, AND RADOMES
	(a) Bartelt, Carl A., Preliminary Investigation, WADC Technical Report No. 52-34, Part 1, "Autogenous Electrification of a B 45A Aircraft". (b) See Par. 4.3.5 (a). (c) See Par. 4.3.5 (b). (d) See Par. 4.3.5 (c).
4.3.8	STATIC DISCHARGER AN/ASA-3
	(a) Handbook, Static Discharger AN/ASA-3. (b) Naval Research Laboratory, "Installation of Dry Wick Type Electrostatic Dischargers", NRL Report No. A903A.
4.3.9	BLOCK AND SQUIRTER DISCHARGERS
	Army-Navy Precipitation Static Report No. A1043A.

Appendix — Title

II DEFINITIONS

See Par. 1.7.1.

IV APPLICATION OF FOURIER INTEGRAL ANALYSIS TO SIMPLE TRAPEZOIDAL PULSES

Campbell, G.A., and Foster, R.M., "Fourier Integrals for Practical Applications", Bell System, B-584, 1931.

V FURTHER CONSIDERATIONS FOR USE OF THE IMPEDANCE CONCEPT

(a) AMC, "Impedance of Aircraft Electrical Systems at Radio Frequencies", Engineering Division Technical Note Serial TN-TSELS6-8, June, 1946.
~~(b) See Par. 1.4 (b).~~
(c) See Par. 3.4.1 (a).

VI THE MUTUAL INDUCTANCE BETWEEN TWO SETS OF INFINITELY LONG PARALLEL PAIRS OF STRAIGHT WIRES

Grover, F.W., "Inductance Calculations", D. Van Nostrand & Co., 1946.

VIII SHIELDED ROOMS - CONSTRUCTION AND USE

See Par. 1.8.2.2 (d).

X MEASUREMENT OF INSERTION LOSS OF FILTERS

(a) Armed Services Electrical Standards Agency, "Method of Insertion-Loss Measurement for Radio-Frequency Filters", Project 116A, Proposed MIL-STD., October, 1951.
(b) MIT, Radiation Lab., "A Method of Shielding for Filter Insertion Loss Measurement", Rad. Lab. Report No. 786, PB 2784.
(c) Signal Corps. "Measurement of Filter Insertion Loss at High and Ultra-High Frequencies", Eng. Lab., Tech. Memo M-1328, September 22, 1950.
(d) USNEL, "Discussion of Problems Relating to Insertion Loss Measurement and Standardization", Report No. 84, Sept. 21, 1948.

XI METHODS OF MEASURING THE EFFECTIVENESS OF SHIELDS

(a) Coles Signal Lab., "Development of Magnetic Field Probe", Tech Memo M-1082, June 10, 1947.
(b) Coles Signal Lab., "Development of Procedure for Testing

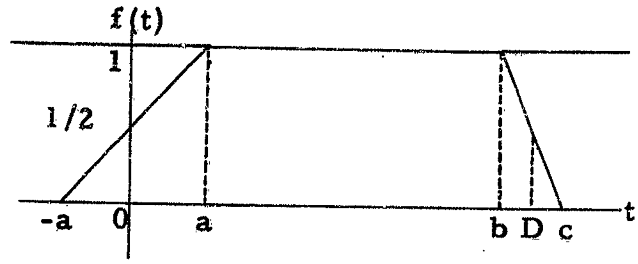
Appendix	Title
	R-F Leakage in Shielding Conduit", (0.15-156 mc), October 21, 1946.
	(c) Schelkunoff, S.A., "The Impedance Concept and Its Application to Problems of Reflection, Refraction, Shielding, and Power Absorption", Bell System Technical Journal, Vol. 17, January, 1938.
	(d) Signal Corps, "Development of Procedure for Testing R-F Leakage in Shielding Conduit (0.15-156 mc)", Tech. Memo No. M-1019, October, 1946, Sig. Corps Engineering Laboratories.
	(e) See Par. 1.8.2.2 (d).
	(f) See Par. 1.8.2.2 (e).
	(g) See Par. 3.1.2 (f).
	(h) See Par. 3.1.2 (g).
XV	THE CONSTRUCTION OF BONDING JUMPERS FOR SHOCK-MOUNTED EQUIPMENT
	Cook Research Laboratories, "Low Impedance Path Bonding Jumpers", Report No. FPRI-1-E.O. No. 112-133, p. 343, June, 1951.
XVI	THEORY OF SHIELDING
	(a) Schelkunoff, S.A., "A Theory of Shielding", Bell Lab. Record 14; p. 229-232, March, 1936.
	(b) Lyons, "Electromagnetic Shielding", Proc. of IRE.
XVII	DECIBEL EQUIVALENTS TO CURRENT, VOLTAGE, AND POWER RATIOS
	Vogelman, J.H., "The Derivation and Application of the Decibel", Eng. Div. Watson Labs., Report No. LESV-2-092447, April 24, 1947.

APPENDIX IV

APPLICATION OF FOURIER INTEGRAL ANALYSIS TO SIMPLE TRAPEZOIDAL PULSES

Let f (t) represent any trapezoidal pulse as follows:

$f(t) = 0;$	$t < -a$
$f(t) = 1/2 + k_1 t;$	$-a < t < a$
$f(t) = 1;$	$a < t < b$
$f(t) = 1/2 - k_2 (t - D);$	$b < t < c$
$f(t) = 0$	$c < t$



The parameters, k_1 and k_2, are the slopes of the rising and falling portions of the pulse, respectively. The points at which f (t) equals 1/2 for t = 0 and t = D are fixed so that the area under the pulse given by the integral of f (t) is a constant independent of the values of k_1 and k_2. Since the total energy of the pulse is proportional to the area, the energy also is independent of k_1 and k_2. The Fourier transform of f (t) is:

$$F(\omega) = \frac{1}{2\pi}\int_{-\infty}^{\infty} f(t)\, e^{-j\omega t}\, dt$$

$$= \frac{1}{2\pi}\left\{ \int \left[\frac{1}{2} + k_1 t\right] e^{-j\omega t}\, dt + \int_a^b e^{-j\omega t}\, dt + \int_b^c \left[\frac{1}{2} - k_2 (t - D)\right] dt \right\} \qquad (1)$$

After carrying out the integrations, and making use of the fact that $a = 1/(2k_1)$, $b = D - 1/(2k_2)$, and $c = D + 1/(2k_2)$, one obtains

$$F(\omega) = \frac{1}{2\pi\omega^2}\left[k_2 \sin \omega D \sin\frac{\omega}{2k_2} + j\left(k_2 \cos \omega D \sin\frac{\omega}{2k_2} - k_1 \sin\frac{\omega}{2k_1}\right)\right] \qquad (2)$$

The quantity of interest is $|F(\omega)|$, the absolute value of the frequency function, which is

$$|F(\omega)| = \frac{1}{2\pi\omega^2}\sqrt{k_2^2 \sin^2\frac{\omega}{2k_2} + k_1^2 \sin^2\frac{\omega}{2k_1} - 2k_1 k_2 \cos \omega D \sin\frac{\omega}{2k_1} \sin\frac{\omega}{2k_2}} \qquad (3)$$

As it stands, this is too complicated to allow an easy interpretation. If cos ωD varies much faster than the other terms, i.e., $D \gg 1/k_1$ and $D \gg 1/k_2$, an envelope

of $|F(\omega)|$ may be established by noting that for $\cos \omega D = \pm 1$,

$$|F(\omega)| = \frac{1}{2\pi\omega^2} \left(k_1 \sin \frac{\omega}{2k_1} \pm k_2 \sin \frac{\omega}{2k_2} \right) \qquad (4)$$

When $\omega \ll k_1$ and $\omega \ll k_2$, $\sin(\omega/2k_1) \simeq \omega/2k_1$ and $\sin(\omega/2k_2) \simeq \omega/2k_2$, so that

$$F(\omega) = \frac{1}{4\pi\omega} (1 \pm 1) \qquad (5)$$

Hence, the low frequency response is independent of k_1 and k_2. At high frequencies, another envelope is obtained by setting $\sin(\omega/2k_1) = \pm 1$ and $\sin(\omega/2k_2) = \pm 1$. Hence,

$$|F(\omega)| = \frac{1}{2\pi\omega^2} (k_1 \pm k_2) \qquad (6)$$

and the high frequency components increase linearly with both k_1 and k_2.

In order to plot $|F(\omega)|$, it is assumed, for simplicity, that $k_1 = k_2 = k$. Then,

$$|F(\omega)| = \frac{1}{\pi\omega^2} k \sin \frac{\omega}{2k} \sin \frac{\omega D}{2} \qquad (7)$$

The curves of Figures 1.2-D and 1.2-E show how this function is affected by a variation of k and D.

APPENDIX V

FURTHER CONSIDERATIONS FOR USE OF IMPEDANCE CONCEPT

The purpose of this appendix is to stimulate thought and promote discussion on the applicability of the "impedance concept" to radio-interference problems. The method of the "impedance concept" is anything but new, but its application to radio-interference problems has lagged due to the lack of sufficient basic data. This lack severely limits the practical application of the method at this time. It is hoped however, that design and research engineers will be stimulated into developing this method sufficiently to make it a true, practical aid in the solution of radio-interference problems. Difficulties that may arise are no argument against the method itself, nor do they detract from its potential fertility. As a first step towards tapping this fertility, this appendix will point out what information will have to be obtained in order to make practical use of the impedance concept, how some of the difficulties in obtaining this information may be overcome, and how this information, if it were obtained, could be used to establish practical methods of reducing radio interference.

The "impedance concept" is the foundation of engineering transmission theory. Authorities in this field have stated that components of transmission systems must have their properties expressed in terms of appropriately chosen impedances, or else a new transmission theory must be developed.

The "impedance concept" was defined in the text as the idea of considering the source of radio interference, the network or medium through which it is transmitted, and the receiver whose effectiveness it finally impairs, as one single entity. The term "impedance" is a general term defined as the ratio of the cause to the effect. In electrical systems, it is the ratio of the electromotive force to the current that flows as a result of that force. The same idea may be applied in other fields. For example, in mechanical systems the impedance is defined as the ratio of the force to the velocity. It is therefore quite common for the design engineer to refer to mechanical impedances and acoustic impedances, as well as to mixed mutual impedances and even the "impedance of free space".

This approach may at first appear "impractical" to some engineers engaged in "interference research". But careful study of the subject brings out the fact that interference problems are basically of the same nature as the transmission problems which, in the fields of telephony, telegraphy, and other types of communication, have been solved successfully by the application of the rigorous analytical methods of circuit analysis. In this process, the source is characterized solely by its generated electromotive force and internal impedance (each assumed to be a known function of frequency over the entire range of interest), the transmitting network or medium by its three image parameters (assuming that the network is linear), and the receiver simply by its impedance. Thus, it takes just four impedances and one image transfer constant to specify such a system completely. This is the justification for the term "impedance concept". The impedances are simply the internal generator impedance, the image impedances looking into the input and output of the transmission network or medium, and the load or receiver impedance. The

transfer constant gives the attenuation and the phase shift of the currents or voltages under consideration.

The definition of the term "impedance" as given above requires further elaboration. For the impedance to have a definite, single value, the cause and the effect of which the ratio is being taken must both vary sinusoidally in time. The case of being constant in time is included as that of varying sinusoidally with zero frequency. The impedance is, in general, a complex quantity, its magnitude being equal to the ratio of the magnitudes of the cause and effect, and its phase angle giving the difference in phase between the two. Next, the definition is extended to quantities that vary periodically, but not necessarily sinusoidally, in time. In this case, use is made of a Fourier series, as explained in Paragraph 1.2. The periodic variation is represented as a sum of an infinite number of sinusoidal terms. With each term there is associated a definite frequency and a definite impedance. The impedance of the system has no longer a single, definite value, but is now a function of frequency, assuming a definite value at each discreet frequency contained in the Fourier series. Finally, the definition is extended to non-periodic quantities. Now a Fourier-integral expansion (see Paragraph 1.2) is required, and the impedance becomes a continuous function of frequency. It should be pointed out that such a Fourier-series or Fourier-integral expansion need not be carried out explicitly. <u>The knowledge that such an expansion is possible suffices to permit the use of the impedance of a system (considered as a continuous, complex function of frequency) in formulating and analyzing any transmission problem.</u>

In most practical transmission problems, the initial datum is the signal to be transmitted. An analysis is then performed of the various types of transmission systems including the terminal equipment at both ends so that the most suitable system may be chosen and a basis for design may be established. The interest usually centers on the proper transmission of the desired signal. Sometimes, however, for example in the design of filters, the attenuation of signals other than the one to be transmitted is a consideration also. Even then, the impedances of the source and the load into which the filter operates usually need to be considered only at the frequencies to be passed. There is the possibility, of course, that the source or load impedances may become very low or high, or go to zero or infinity, at some important frequencies to be attenuated, and therefore, the designer should have some idea of the frequency characteristics at all frequencies in order to avoid the possibility of sharp dips in the attenuation characteristics of the filter or network being designed.

The application of the impedance concept, which has proved so fruitful in closely related problems, to the radio interference problem will prove increasingly profitable as more pertinent data, on which an analysis can be based, become available. In fact, many engineers working in this field are already taking advantage of considerable data now available and are improving their measuring equipments and techniques to acquire additional data on which to base more efficient designs. Difficulties faced by the interference engineer due to insufficient impedance data and the general unavailability of conveniently usable field-type instruments tend to set a limit to the applicability of the impedance concept to practical design problems at this time. But it is hoped that forward-looking engineers will take advantage of this method to an increasing degree as more adequate measuring techniques and instruments are developed and more pertinent data becomes available.

The impedance point of view is less helpful in the modification of existing equipments and installations (i.e., "field fixes") than in the incorporation of interference-reducing practices into the original designs. For, once a particular piece of equipment has been designed and manufactured, its impedance at all frequencies is fixed and very difficult to change. If the design was carried through without giving any consideration whatever to the impedances at the radio frequencies likely to be encountered, then one cannot expect that by slight changes in the transmitting network - such as the addition of capacitors or filters, or the rerouting of wiring - an efficient overall system is produced having all the desired and none of the undesired properties. To produce such a system, the impedance point of view must be adopted from the very beginning and must be taken into consideration in the design of terminal equipment at both ends as well as of the interposed transmission network. Nevertheless, this does not mean that the existing equipment is useless from the impedance point of view. No design can proceed without a wealth of pertinent basic data. Measurements on existing equipments can contribute a great deal to such data, and in fact, until other and better equipments become available, existing equipments are the only sources from which such data can be obtained.

The radio interference problem can be formulated in the following way: Given a piece of apparatus - say a direct current motor - which is known to be a potential generator of interfering voltages or currents, given a receiver which may be affected by these interfering currents or voltages, and given finally a transmitting network, which is really the entire aircraft with all its electric and electronic equipment and associated wiring. The problem is to find the dependence of the six parameters of this system on all variables under the control of the designer. Once this dependence is known, transmission theory can be used to determine the optimum values of the system parameters, and the design can proceed to obtain the proper values of the controllable variables.

If impedances are to be measured in the conventional ~~manner, terminals~~ must be designated between which the measurements are to be made. At very high and ultra-high frequencies such terminals may not be available, and the very word "impedance" must be redefined to acquire a meaning applicable to such situations. This introduces a new type of difficulty, which will be discussed later. For the time being it will be assumed that the frequencies are low enough to insure meaningfulness of the expression "impedance between a set of two terminals".

After the statement of the problem, the first step is to choose two pairs of terminals, one to be the junction between the source and the transmitting network, and the other one to be the junction between the transmitting network and the receiver. Since the interfering signal may leave the source by any one of several routes, and may enter the receiver at any one of several places, this choice itself is not an easy one to make and involves certain simplifications at the outset. If the source is completely surrounded by an effective shield so that radiated interference may be neglected and conduction through the power lead is the only way by which interfering signals can leave, then the approximation involved in considering only one pair of terminals at the source is a very good one. Similarly, if the receiver is adequately shielded and employs a shielded lead-in from the antenna, and if direct coupling to the antenna is negligible, so that again the power lead is the only path of entry, then all terminals but two at the receiver may safely be neglected. In other cases these approximations may be quite poor. At any rate, as in all complicated

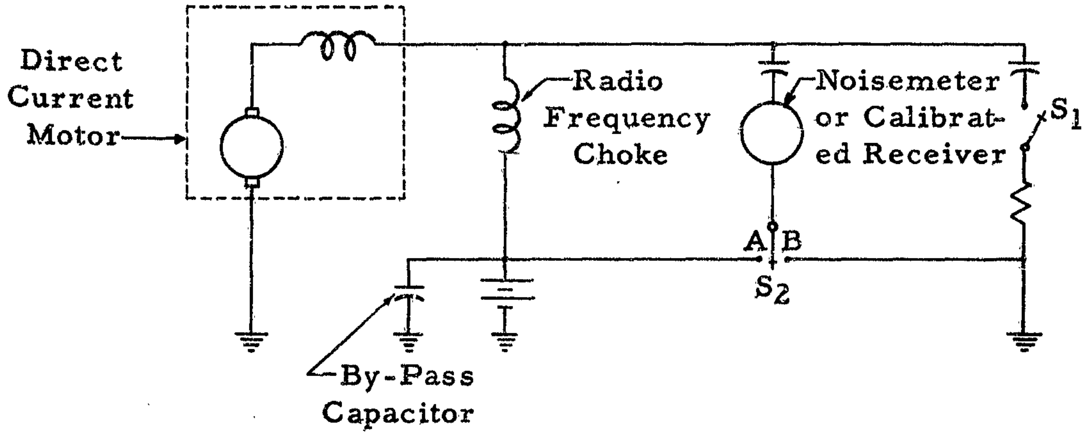


Fig. V-2. Measurement of Open-Circuit Voltage and Short-Circuit Current Using Noisemeter or Calibrated Receiver

of the generated electromotive force and the absolute value and phase angle of the internal impedance of the source as functions of frequencies may be prepared from these data.

The above method of measuring the short-circuit current is applicable only when the interference voltages are comparatively high. It may happen that the short-circuit current is too small to produce a measurable indication on the meter. Then an alternate method may be employed using a signal generator with calibrated output and known internal impedance as shown in Figure V-3. The signal generator is connected in parallel with the battery, a large series capacitor being used to protect the generator from the direct current. The output of the signal generator must be much larger than the interference voltages generated by the source so that the latter can be neglected. From a knowledge of the voltage as read on the signal generator, its internal impedance (usually a pure resistance), and the current, the absolute value of the impedance may again be determined. The reactance may again be found by connecting a standard inductance or capacitance in series and resonating it with the reactance of the source. This second method is not subject to frequency and wave-form errors provided that the output of the calibrated signal generator is a pure sine wave.

Similar measurements must also be made at the input of the receiver. Again the receiver must operate normally if the measurements are to be meaningful. However, since the receiver is a passive network, bridge methods may be used provided precautions are taken to prevent the power currents (normally at 400 cycles per second in aircraft) from damaging the bridge or affecting the measurements. A possible circuit is shown in Figure V-4. Alternately a signal generator and milliammeter may be used in the circuit of Figure V-5.

To avoid the complexities of the impedance measurements just described, the impedance of the source and the receiver may be measured without supplying any power. Recent tests have shown that fair results may be expected using this method, and data so obtained may well serve as a good first approximation.

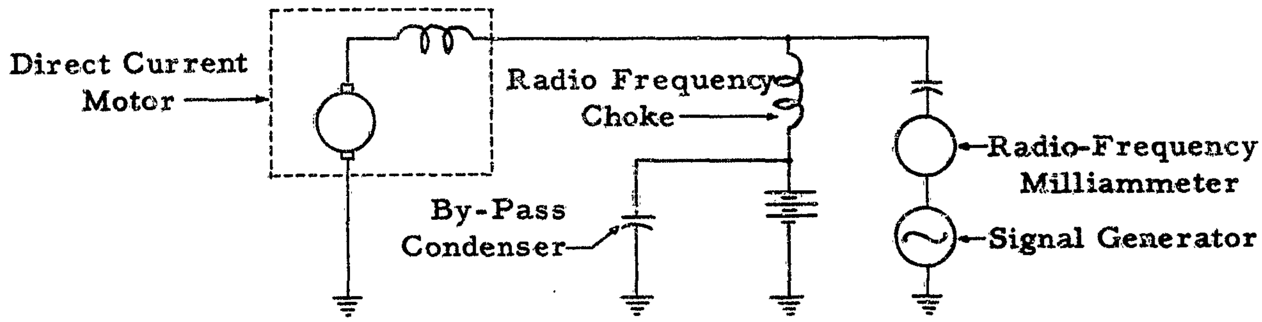


Fig. V-3. Measurement of Internal Impedance with External Signal Generator.

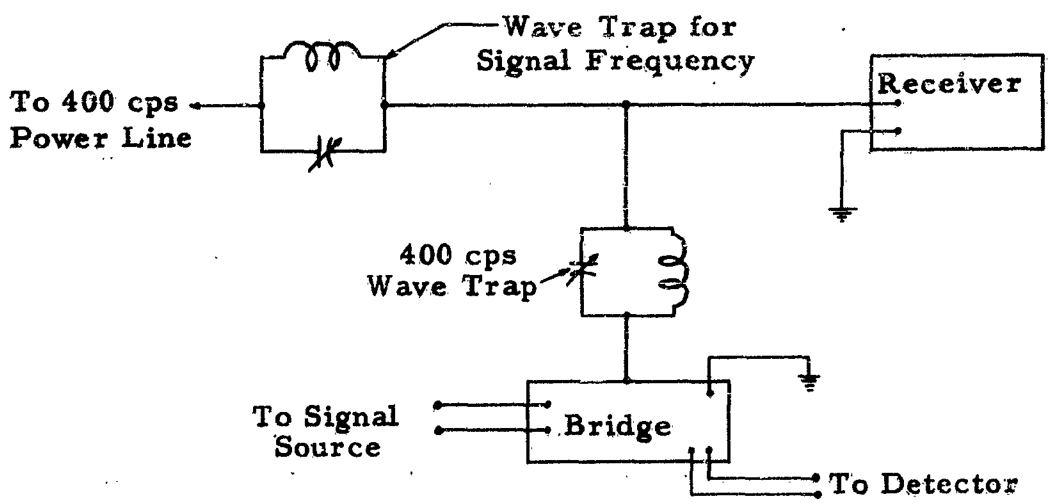


Fig. V-4. Bridge Measurement of Receiver Input Impedance

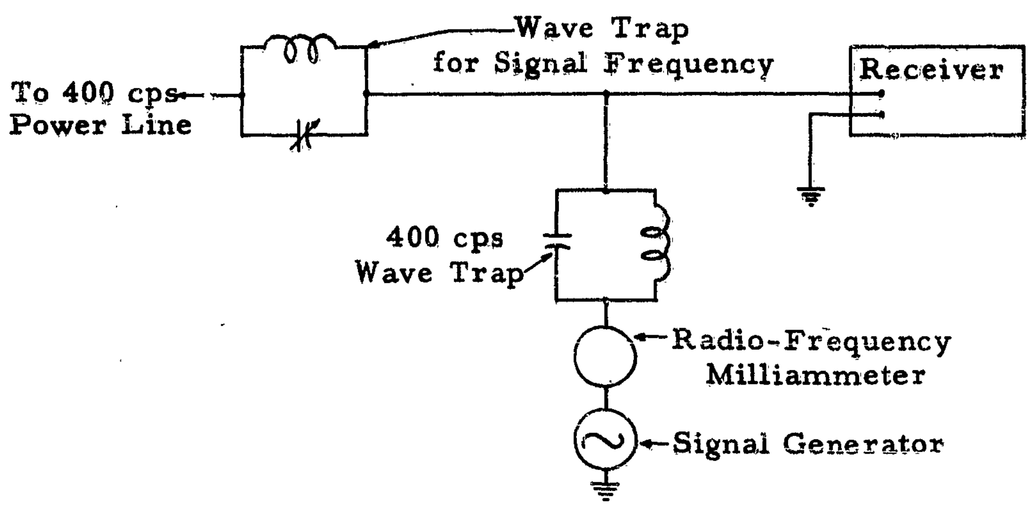


Fig. V-5. Alternate Method of Measuring Receiver Input Impedance

Finally, the parameters of the transmission network must be measured. These may be obtained from measurements of the input impedances at each end with the other end open-circuited and with the other end short-circuited. If the open- and short-circuit impedances at the input are called Z_{OC} and Z_{SC}, respectively, and the corresponding impedances at the output are called Z'_{OC} and Z'_{SC}, then the image impedances Z_{I1}, and Z_{I2}, and the image transfer constant, θ, are obtained from:

$$Z_{I1} = \sqrt{Z_{oc} Z_{sc}} \quad (1)$$

$$Z_{I2} = \sqrt{Z'_{oc} Z'_{sc}} \quad (2)$$

$$\theta = \tanh^{-1} \sqrt{\frac{Z_{sc}}{Z_{oc}}} = \tanh^{-1} \sqrt{\frac{Z'_{sc}}{Z'_{oc}}} \quad (3)$$

The open- and short-circuit impedances may be measured either by bridge methods or by means of a signal generator and milliammeter as before.

After measuring methods for these six quantities have been established, a large number of sets of measurements must be taken to determine the effect of the variation of various controllable design quantities such as size of wires, position of ground leads, shape of frames, brush materials, and many others. This is a huge task, yet one that is necessary if practical results are expected. Only if the effect of each design detail is known can one hope to be able to control the impedances and generated voltages by proper design before the equipment is actually built.

To be of practical use to the designer, the information obtained must be presented in the form of charts or curves. For example, one such curve would give information about the variation of the image impedances of the aircraft wiring with the size of wires used. Another might present data on the dependence of generated interference voltages on brush pressure in a motor. (See Figure 3.2.1.1.1-A). A third would give the input impedance of a receiver (at the power leads) as a function of ground lead location. It is quite obvious that the number of charts and curves required is very large.

Measurements of the type just described have never been made in actual installations on an extensive scale. Isolated attempts are on record of obtaining significant data in a few special cases. Not many conclusions can be drawn from the results of these attempts, yet they bring out clearly the difficulties as well as the definite possibility of overcoming these difficulties. In addition, certain observations are of interest even though they require thorough checking in view of the special conditions under which they were made. These observations are enumerated below:

(a) Measurements of generated interference voltages made by different methods are difficult to correlate. If made according to Specification JAN-I-225, they will yield consistent results and thus be suitable for comparisons and standardization, but the true generated electromotive forces cannot be obtained in this way because measurements are made with a load. Figures V-6 and V-7 show the interference voltages of three motors at various frequencies, the first figure showing the results obtained according to JAN-I-225, the second those obtained when the voltages were measured across a load consisting of six feet of special high-loss coaxial cable. The second figure probably gives more nearly the true generated electromotive forces because the cable presented a high resistive impedance at all interference frequencies.

(b) The generated electromotive forces decrease very rapidly for frequencies above one megacycle per second for most motors. Most of their energy is concentrated at the lower frequencies.

(c) The impedance of the aircraft wiring is such that the frequencies above one megacycle per second are accentuated by the resonances of the wiring system. This results in several high peaks of voltages or currents, their number depending on the length and the complexity of the wiring.

(d) For the frequencies of interest (from 0.15 to 1000 mc) the actual attenuation of the wiring system is very small. Apparent attenuation in any one wire may be large because of the spreading of the interference energy over the entire wiring system.

(e) Transmission and spreading of the interference energy is almost entirely by conduction and inductive and capacitive coupling. Transmission by radiation is negligible below 150 megacycles.

(f) Measurements of the impedance of the wiring system show wide and rapid variations of both the resistive and reactive components with frequency. Observed maxima and minima bear no apparent harmonic relationship.

(g) Serious impedance mismatches are present at the junction points of the wiring system and at the points where the wiring is terminated in electrical or electronic equipment. These mismatches lead to reflections and standing waves on the wires resulting, in turn, in heavy concentrations of interference energy in certain regions of the system. Broad-band matching would eliminate these concentrations, but would also deliver more of the energy to the terminal equipment. At any rate, an attempt must be made to dissipate the energy before it can do any harm rather than simply redistribute it in space or in the frequency spectrum.

No information is available on the effect of controllable factors on the measured parameters. Even if such information were available, an investigation must first be made into which values of the parameters are most desirable from the point of view of eliminating radio interference.

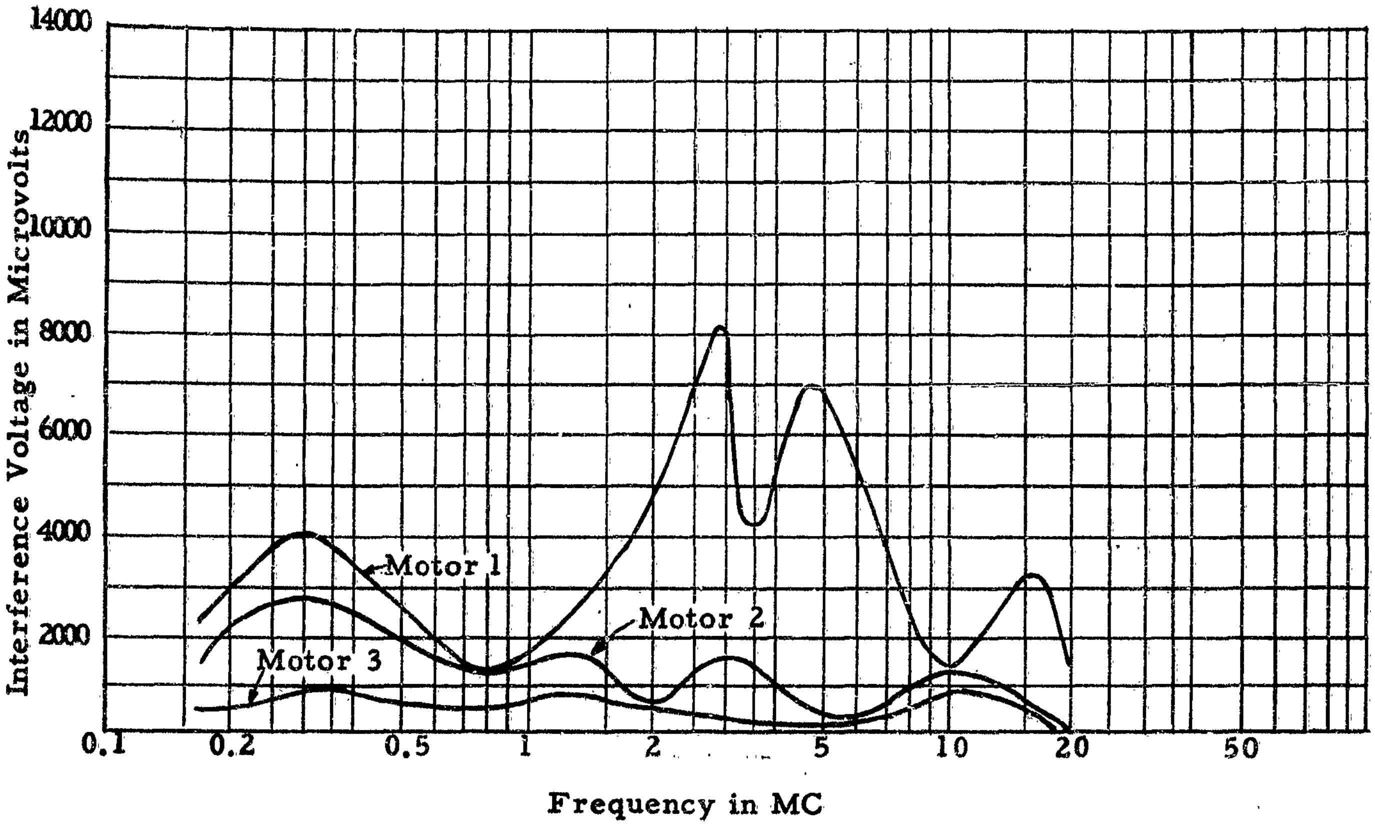


Fig. V-6 Conducted Radio Interference of Three Motors
Measured According to Spec. JAN-I-225

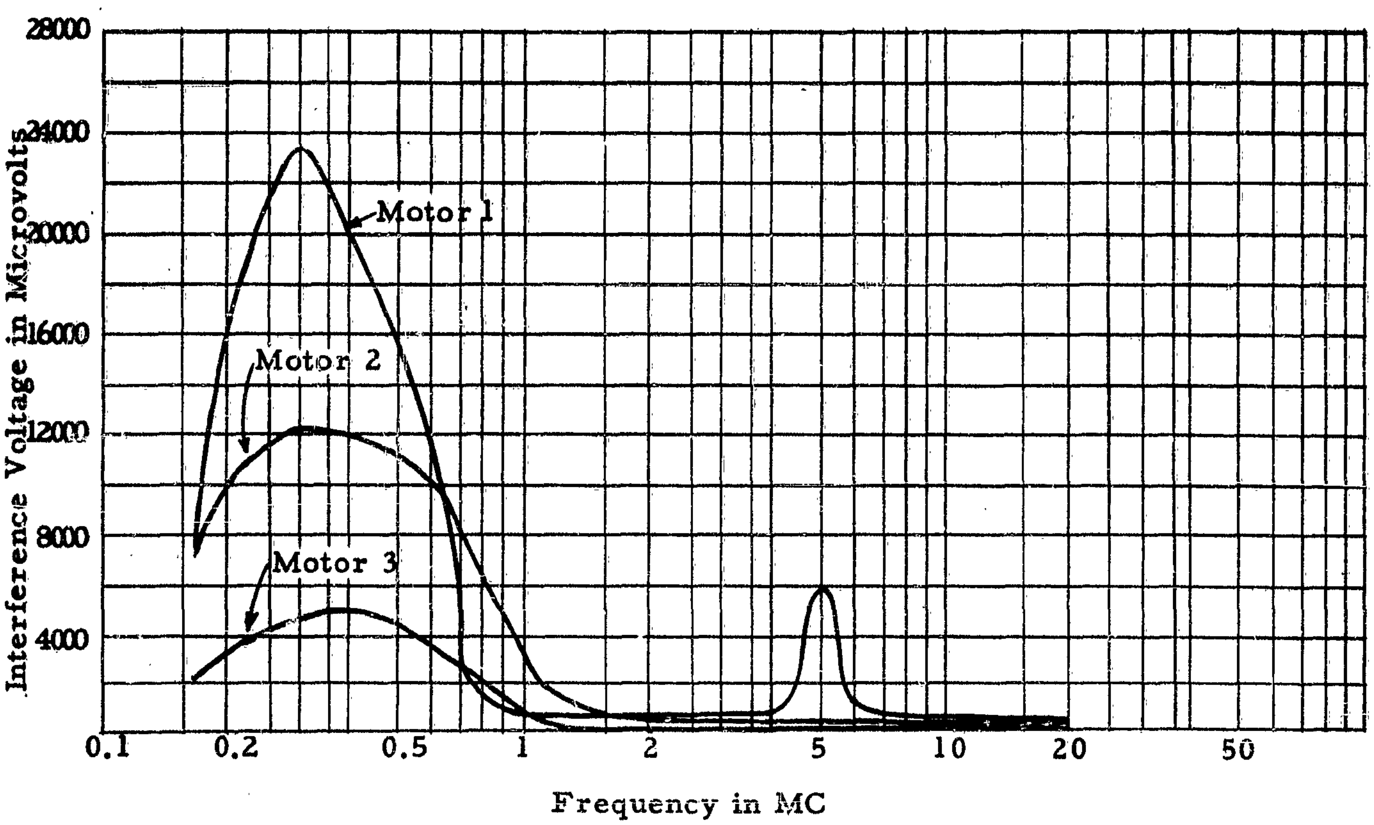


Fig. V-7 Conducted Radio Interference of Three Motors
Measured Through High-Loss Six-Foot Coaxial Line

Ideally, the requirement is that no interfering currents or voltages shall reach the receiver. Assuming that the generation of interference electromotive forces cannot be prevented (though the measurements suggested here would form a basis for designing the equipment so that their generation is minimized), the problem then is to attenuate the interference as much as possible. As pointed out in Paragraph 1.4, a signal undergoes modification in four distinct ways as it is transmitted from a source to a receiver. There is the attenuation proper in the transmitting network, there are the reflection losses at the input and output of the transmitting network, and there is the interaction loss due to multiple reflections. Of these, only the first, the attenuation proper, is always a true attenuation (or zero). Any one of the other three "losses" may be negative, and hence may be a gain rather than a loss. If the observations enumerated above are trustworthy, then the attenuation proper cannot be relied upon to provide a substantial reduction of interference (see item d above). Therefore, the reflection and interaction losses must be made to produce the necessary reduction unless it is possible to redesign the wiring system so that true attenuation is obtained within the transmitting system. (One suggestion to achieve this has been to use lossy cables for all the major wiring in the aircraft).

Figure V-8 shows how the reflection loss varies with the magnitude and phase angle of the "mismatch ratio", i.e., the ratio of the impedances looking to the right and to the left of a pair of terminals. This indicates that this loss is positive for all magnitudes of the ratio only when the phase angle, φ is zero. For all other phase angles, the loss is negative for a portion of the range plotted, and this portion as well as the magnitude of the gain both increase substantially as the phase angle

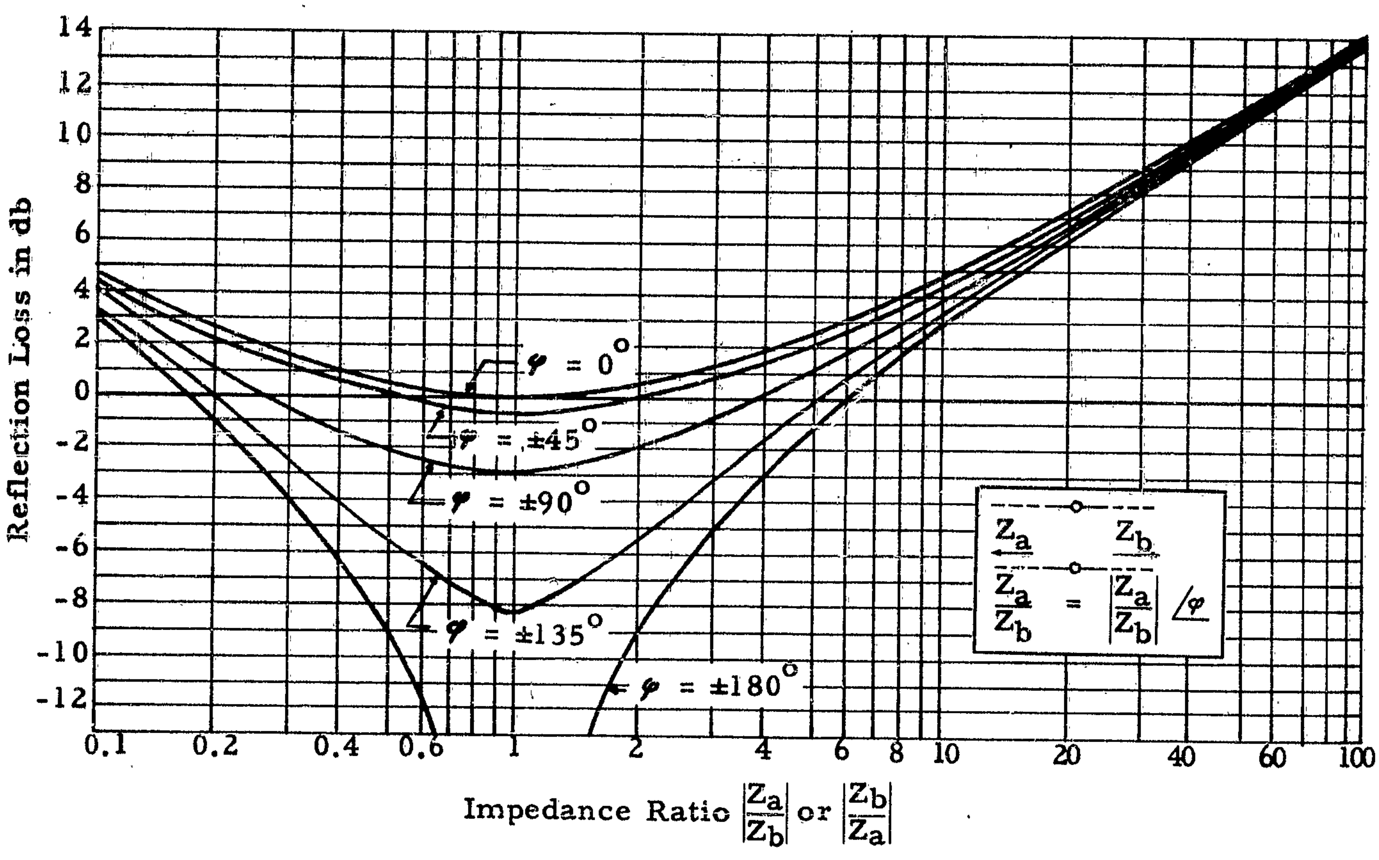


Fig. V-8 Reflection Loss Resulting from Mismatch of Impedances at a Junction

varies towards 180° either through positive or through negative values. This figure would have to be used carefully in determining what impedances would be desirable for maximum attenuation. A similar chart should be worked out for the interaction loss.

If the properties of the system itself cannot provide sufficient attenuation, a filter could be inserted between the output terminals of the source and the input terminals of the transmitting network. The most efficient filter providing the maximum attenuation with the minimum expenditure of weight and space can be designed only when the impedances between which it is to operate as well as the amplitudes and frequencies of the currents it is to transmit and attenuate are precisely known. The weight and space requirements for most radio interference filters could be reduced substantially without sacrificing effectiveness if full information about the impedances were available. In fact, this information would not even be required over the entire frequency range from 0.15 to 1000 mc, but only for the lower frequencies from 0.15 to about 3 or 4 mc because the generated interference has by far the largest amplitudes in this lower region. The drawback of trying to utilize this saving in weight and space is that a filter so designed would be effective only in the particular position and for the particular system for which it was designed. In the absence of more detailed information, the slightest change in design, such as a change in the size of wires or in the material of the brushes, or even the slightest change in location of wiring or equipment, will have a large and entirely unpredictable effect on the impedances of the system, and hence may make the filter ineffective. Until ways are found of controlling the impedances and predicting their changes with changes in design parameters, filters must obviously be designed so as to remain effective under a large variety of conditions.

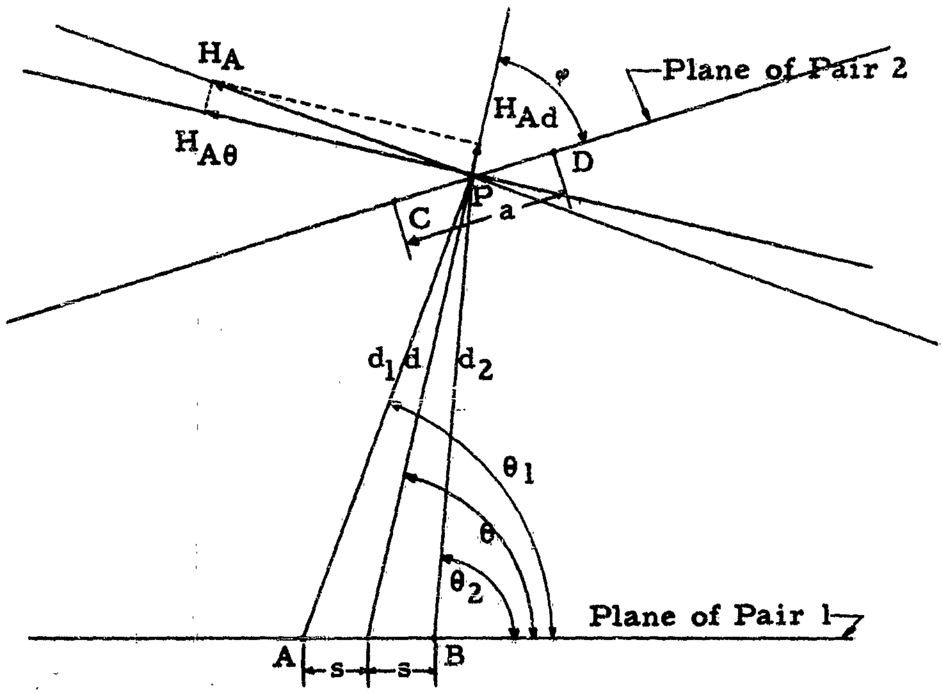
If this general method is to be extended to very high and ultra-high frequencies, additional difficulties must be overcome. Above 100 megacycles, radiation will become important, and if the circuit concept is still to be used, the radiation resistance must be included in the impedances. In certain cases, the term "impedance" itself must be given a new meaning. Instead of dealing with the impedance of a network, defined as the ratio of voltage to current, it may be necessary to deal with the impedance of a transmitting medium, which is defined as the ratio of the electric to the magnetic field intensity. In defining this ratio, the type of wave that gives rise to the fields, i.e., its geometrical configuration and polarization, must be specified in addition to the properties of the medium itself. Since several types of electromagnetic waves may be present simultaneously, the additional complexity introduced is clearly evident.

In addition to these basic difficulties, practical difficulties are added due to the limited frequency range of conventional measuring instruments. It is true that some voltmeters and one admittance bridge are available that can be used up to 1000 megacycles. But they require special terminations, which usually are not available in aircraft. Impedance determinations at these frequencies are usually made by standing wave measurements. The adaptation of this method to the purpose at hand will require a considerable amount of research and development.

APPENDIX VI

THE MUTUAL INDUCTANCE BETWEEN TWO SETS OF INFINITELY LONG PARALLEL PAIRS OF STRAIGHT WIRES

Consider the cross-section of the four wires A, B, C, and D shown below.



Assume that d >> s, so that

$$d_1 \simeq d + s \cos \theta \tag{1}$$

$$d_2 \simeq d - s \cos \theta \tag{2}$$

Let the left wire of Pair 1 carry a current I out of the plane of the paper and the right wire the current I into the plane of the paper. The two components of the magnetic field H (at Point P), produced by the current in wire A of Pair 1, are then

$$H_{A\theta} = \frac{I \cos (\theta_1 - \theta)}{2 (d + s \cos \theta)} \tag{3}$$

$$H_{Ad} = \frac{I \sin (\theta_1 - \theta)}{2 (d + s \cos \theta)} \tag{4}$$

In the same way, for the field H_B set up by wire B of pair 1,

$$H_{B\theta} = \frac{I \cos (\theta - \theta_2)}{2 (d - s \cos \theta)} \tag{5}$$

$$H_{Bd} = \frac{I \sin(\theta - \theta_2)}{2(d - s\cos\theta)} \tag{6}$$

The total field at P is then

$$H_\theta = H_{A\theta} + H_{B\theta} \simeq \frac{I}{2\pi}\left[\frac{-2\,s\cos\theta}{d^2 - s^2\cos^2\theta}\right] \simeq \frac{-I\,s\cos\theta}{\pi d^2} \tag{7}$$

because $\cos(\theta - \theta_1) \simeq \cos(\theta - \theta_2) \simeq 1$ and $d \gg s$, and

$$H_d = H_{Ad} + H_{Bd} \simeq \frac{I}{2\pi}\left[\frac{2d\sin(\theta_1 - \theta)}{d^2 - s^2\cos^2\theta}\right] \simeq \frac{I\,s\sin\theta}{\pi d^2} \tag{8}$$

because $\sin(\theta_1 - \theta) \simeq \sin(\theta - \theta_2) \simeq \frac{s\sin\theta}{d}$.

The flux density, B, at Point P is $4\pi \times 10^{-7}$ H, and its component perpendicular to the plane of Pair 2 is

$$B_P = 4\pi \times 10^{-7}\,(H_d \sin\varphi + H_\theta \cos\varphi) \tag{9}$$

$$= \frac{4Is}{d^2} \times 10^{-7}\,(\sin\theta\sin\varphi - \cos\theta\cos\varphi) \tag{10}$$

$$|B_P| = \frac{4Is}{d^2} \times 10^{-7}\cos(\theta + \varphi) \tag{11}$$

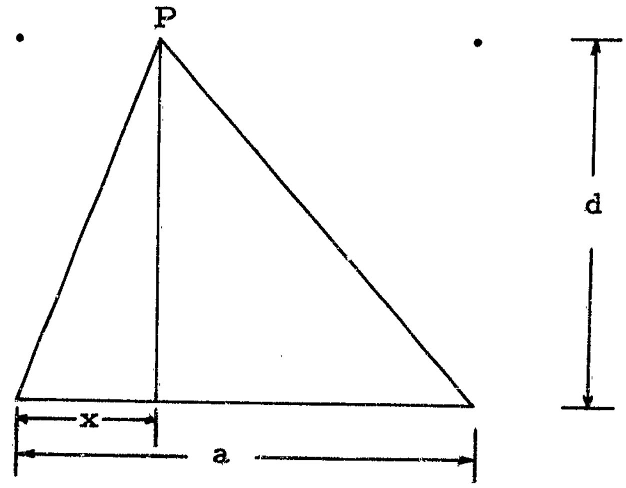
Assuming that $a \ll d$, the flux density B is practically constant between the wires C and D of Pair 2. Hence, the total flux for a unit length of wire is

$$\Phi = B_P\,a = \frac{4Isa}{d^2} \times 10^{-7}\cos(\theta + \varphi) \tag{12}$$

and the mutual inductance (henries/meter) is

$$M = \frac{\Phi}{I} = \frac{4sa}{d^2} \times 10^{-7}\cos(\theta + \varphi) \tag{13}$$

This is a maximum when $\theta + \varphi = 0$, or $\theta + \varphi = 180^\circ$, and zero when $\theta + \varphi = 90^\circ$.



Now consider the above diagram. Here $\theta = \varphi = 90^{o}$ and $2s = a$, for simplicity ; but it is no longer assumed that $d \gg a$. It is assumed, however, that the cross-section of the wires is small as compared with the distances between them. Then, at P,

$$H_P = \frac{I}{2\pi}\left[\frac{x}{x^2 + d^2} + \frac{a - x}{(a - x)^2 + d^2}\right] \tag{14}$$

$$B_P = 4\pi \times 10^{-7} H_P \tag{15}$$

$$\Phi = \int_0^a B_P \, dx = 4\pi \times 10^{-7} \int_0^a H_P \, dx \tag{16}$$

$$= 2I \times 10^{-7} \log\left(1 + \frac{a^2}{d^2}\right) \tag{17}$$

$$M = \frac{\Phi}{I} = 2 \times 10^{-7} \log\left(1 + \frac{a^2}{d^2}\right) \tag{18}$$

If $d \gg a$, this expression for M reduces properly to that on the previous page since, for $x \ll 1$,

$$\log(1 + x) \simeq x \tag{19}$$

APPENDIX VII

LOW PASS FILTER DESIGN DATA

Interference filters may be constructed from reactances in various possible configurations, however, the information presented in this appendix will be confined to ladder filters of the low pass type shown in Figure VII-1.

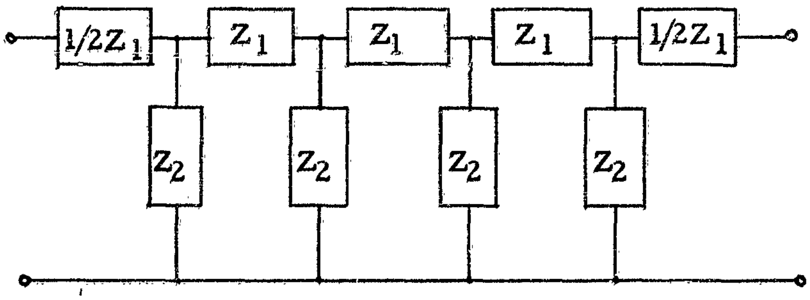


Fig. VII-1 Ladder Network Filter

The series impedance Z_1, and the shunt impedance Z_2 consist of inductive or capacitive reactances or a combination of both.

Both the T and pi-sections, as shown in Figure VII-2, are grouped under the common heading of ladder networks. Each of the two series arms of the T-section is equal to $Z_1/2$ resulting in a full series impedance of Z_1, while each of the two shunt arms of the pi-section is equal to $2Z_2$ resulting in a full shunt impedance of Z_2.

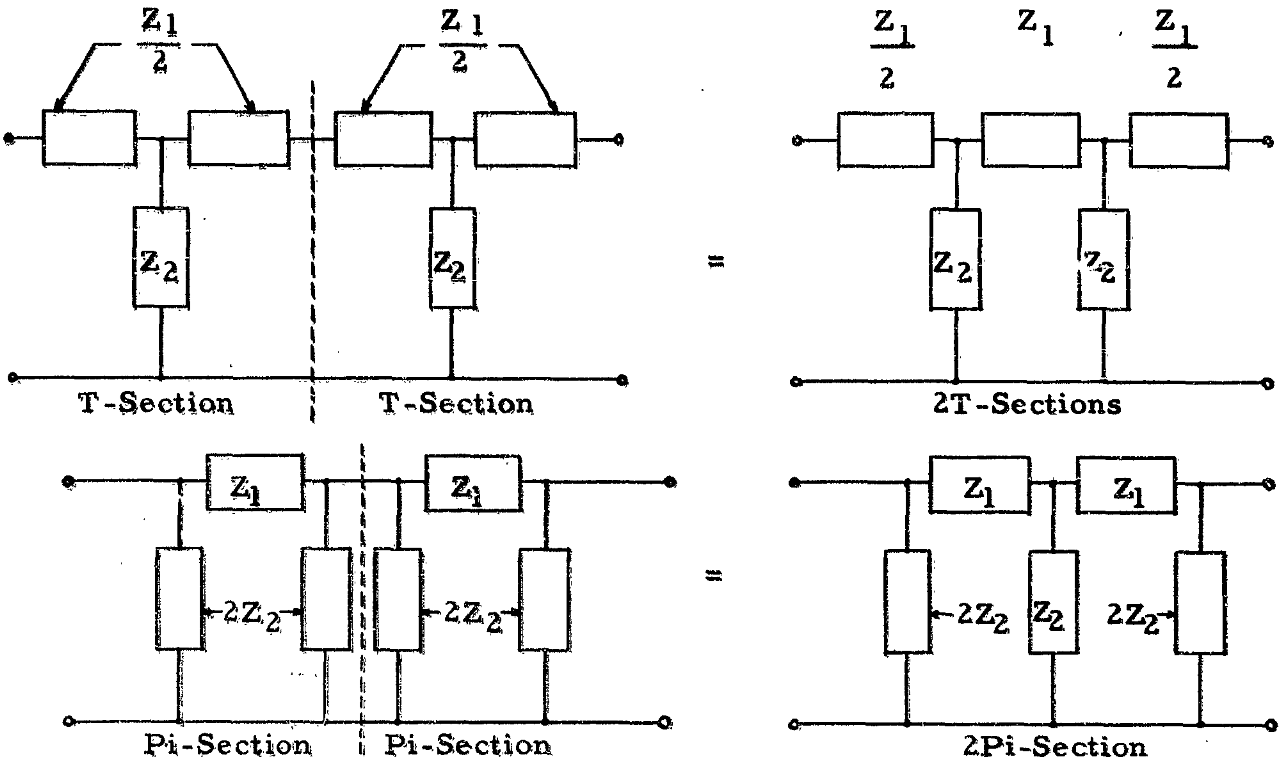


Fig. VII-2 Basic Configurations Included in Ladder Filters

Figure VII-3 shows a T network with three arms Z_a, Z_b, and Z_c, connected to a source of voltage E whose internal impedance is Z_{I_1}, and terminated by an impedance Z_{I_2}. If the impedance Z_{I_1} is equal to the impedance looking in from the terminals 1-2, and similarly, if the impedance Z_{I_2} is equal to the impedance looking in from terminals 3-4, then the impedances Z_{I_1}, and Z_{I_2} are called the image impedances of the T network. A similar relationship can also exist for a pi network.

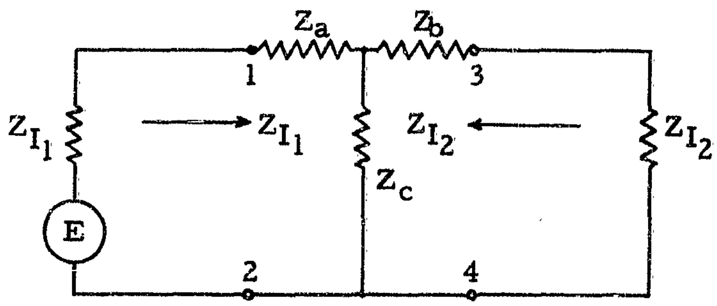


Fig. VII-3 A T Network Terminated by Its Image Impedances

In terms of open-circuit and short-circuit measurement the image impedances Z_{I_1} and Z_{I_2} for both T and pi networks are expressed as

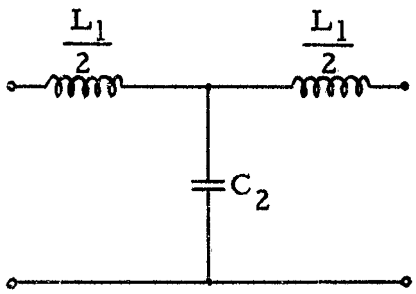
$$Z_{I_1} = \sqrt{Z_{oc}\ Z_{sc}} \tag{1}$$

$$Z_{I_2} = \sqrt{Z'_{oc}\ Z'_{sc}} \tag{2}$$

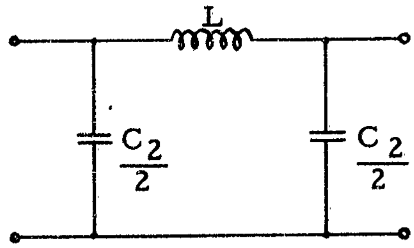
where Z_{oc} and Z_{sc} are the impedances looking into terminals 1-2 with terminals 3-4 open-circuited and short-circuited respectively, and Z'_{oc} and Z'_{sc} are the impedance looking into terminals 3 - 4 with terminals 1 - 2 open-circuited and short-circuited, respectively. When the two image impedances are equal, $Z_a = Z_b$, the filter is symmetrical, and the impedances are equal to Z_o, the characteristic impedance of the network.

Figure VII-4 gives the schematic diagram of two constant-k filters, the equation for their elements, and curves of attenuation plotted as a function of frequency. The filters are so-called because the product $Z_1 Z_2$ is a constant for all frequencies and is equal to k^2. The value k is equal to the value R used in the table of Figure VII-5.

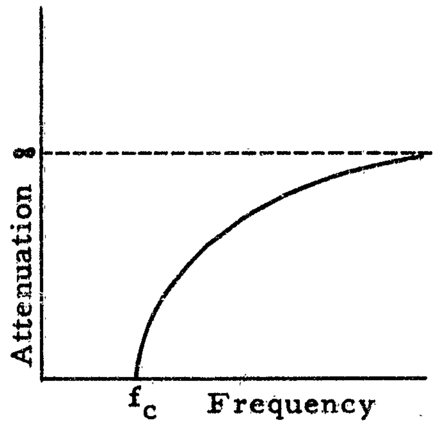
The table given in Figure VII-5 lists the element values for the full series and shunt arms of a low pass constant-k filter whose image impedances are equal to 50 ohms. The following example illustrates the method by which values of L_1 and C_2 for cut-off frequencies not listed in the table may be obtained. A low pass filter working out of and into a 50-ohm line having a cut-off frequency of 150 cycles requires an inductance of 0.106 henries and a capacitance of 42.4 µf. For a cut-off frequency of 1.5 mc both of these values must be divided by 10^4, the ratio of the desired frequency to the listed frequency, because the product L_1C_2 varies inversely with the cut-off frequency.

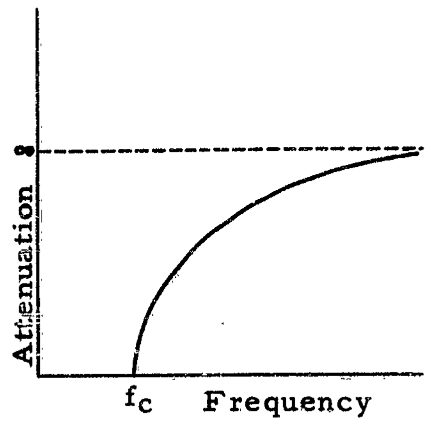


T Type Low Pass Filter



π Type Low Pass Filter





$$f_c = \frac{1}{\pi\sqrt{L_1 C_2}}$$

$$R = \sqrt{\frac{L_1}{C_2}}$$

$$L_1 = \frac{R}{\pi f_c}$$

$$C_2 = \frac{1}{\pi f_c R}$$

$$f_c = \frac{1}{\pi\sqrt{L_1 C_2}}$$

$$R = \sqrt{\frac{L_1}{C_2}}$$

$$L_1 = \frac{R}{\pi f_c}$$

$$C_2 = \frac{1}{\pi f_c R}$$

Fig. VII-4 Conventional Constant-k Filters

The expression $R = \sqrt{L_1/C_2}$ suggests a method by which data for a filter whose characteristic impedance is R, different from the value given in the table, can be obtained. First find the required data for a 50-ohm filter then multiply the inductance by R/50 and divide the capacitance by the same value.

Constant-k type filters act as a resistive load throughout the pass band if they are properly terminated in their image impedance. However, at the cut-off frequency the load becomes zero for the T network and infinite for the pi network as shown in Figure VII-6. At frequencies beyond cut-off the load becomes imaginary. That is, in the attenuation band the filter acts as a reactive load, does not take energy from

the interference source, and, therefore, does not transmit energy to the terminating impedance. However, infinite attenuation of the interference frequencies is obtained only by a filter with purely reactive arms which, of course, exists only in theory.

f_c (cps)	L_1 (mh)	C_2 (μf)
30	5.31×10^2	2.12×10^2
100	1.59×10^2	6.37×10
150	1.06×10^2	4.24×10
200	7.96×10	3.18×10
250	6.37×10	2.55×10
300	5.31×10	2.12×10
350	4.55×10	1.82×10
400	3.98×10	1.59×10
450	3.54×10	1.41×10
500	3.18×10	1.27×10
550	2.89×10	1.16×10
600	2.65×10	1.06×10
650	2.45×10	9.79
700	2.27×10	9.09
750	2.12×10	8.49
800	1.99×10	7.96
850	1.87×10	7.49
900	1.77×10	7.07
950	1.68×10	6.70
1×10^3	1.59×10	6.37
3×10^3	5.31	2.12
10×10^3	1.59	6.37×10^{-1}
30×10^3	5.31×10^{-1}	2.12×10^{-1}
100×10^3	1.59×10^{-1}	6.37×10^{-2}
300×10^3	5.31×10^{-2}	2.12×10^{-2}
1×10^6	1.59×10^{-2}	6.37×10^{-3}
3×10^6	5.31×10^{-3}	2.12×10^{-3}
10×10^6	1.59×10^{-3}	6.37×10^{-4}
30×10^6	5.31×10^{-4}	2.12×10^{-4}

Fig. VII-5 Constant-k Low Pass Filter (R = 50Ω) - Table-1

If a sharper cut-off (higher attenuation in the region beyond the cut-off frequency) than that exhibited by a constant-k type section is desired, it can be obtained by adding additional impedances to the prototype, the constant-k section. When the values of the added impedances are derived from those of the prototype, the resultant section is called an M-Derived filter. These impedances sharpen the cut-off of the section by providing an attenuation which approaches infinity at a frequency beyond cut-off as shown in Figure VII-7.

The position of the added impedance elements in the filter network determine the specific nomenclature of the section. If the additional impedances are added to the series arm of the section, the section is shunt derived. The section is series

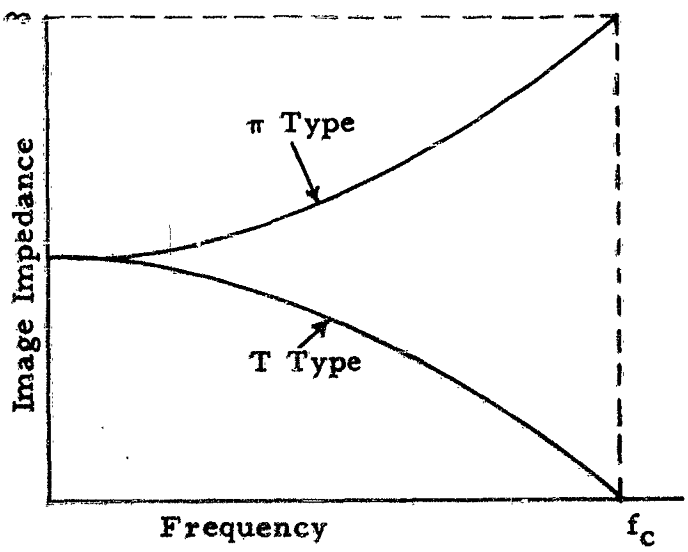


Fig. VII-6 Variation of Image Impedance with Frequency in Low Pass Constant-k T and Pi Type Filters

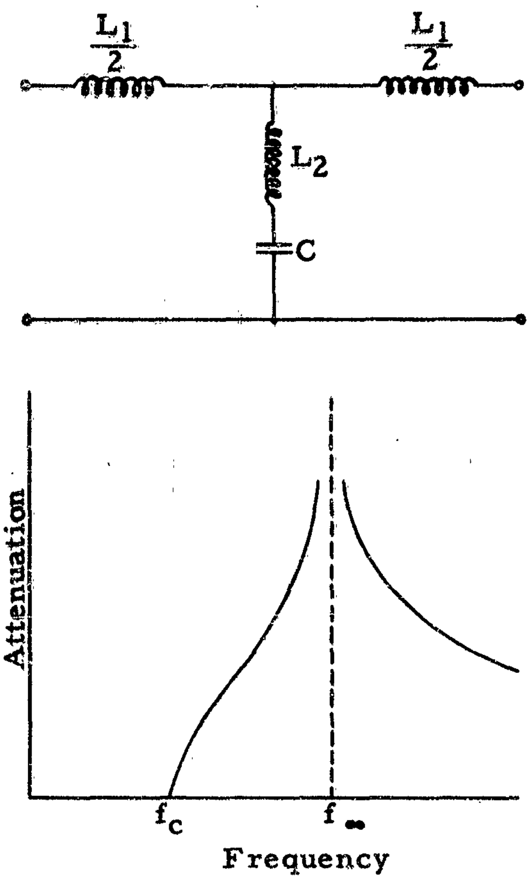
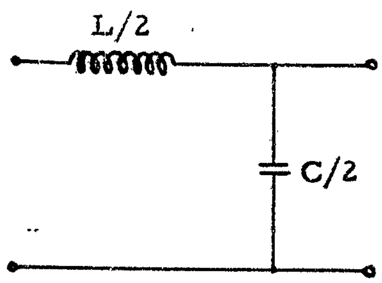
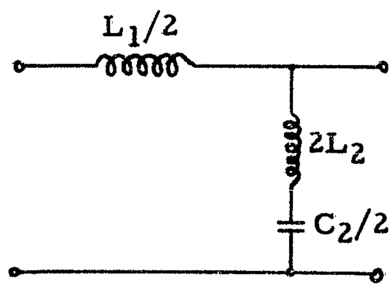


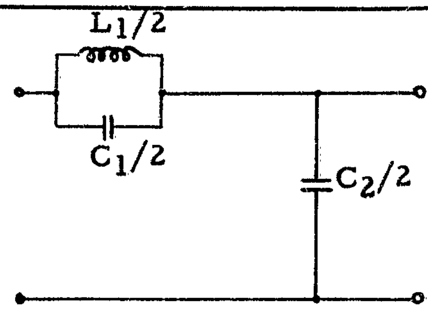
Fig. VII-7 Variation of Attenuation with Frequency in a Low Pass M-Derived T Type Filter



Constant k Half Section



Series m-Derived Half Section



Shunt m-Derived Half Section

$$L = \frac{R}{\pi f_c}$$

$$C = \frac{1}{\pi f_c R}$$

$$m = \sqrt{1 - \left(\frac{f_c}{f_\infty}\right)^2}$$

$$L_1 = mL$$

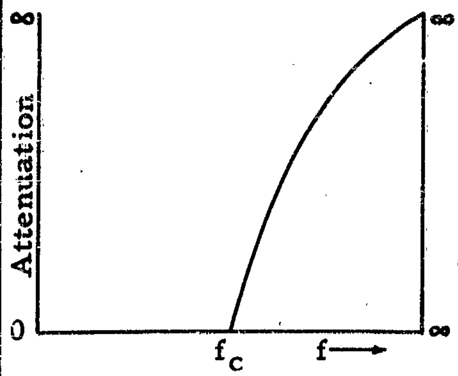
$$L_2 = \frac{1-m^2}{4m} L$$

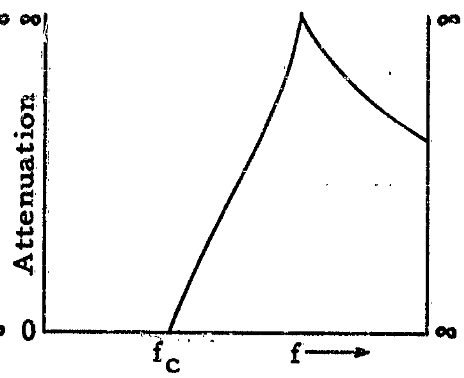
$$C_2 = mC$$

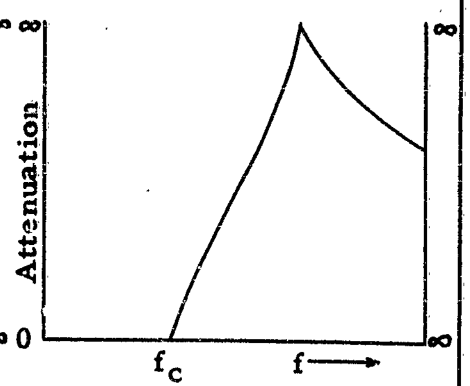
$$L_1 = mL$$

$$C_1 = \frac{1-m^2}{4m} C$$

$$C_2 = mC$$







Note: L in Henrys
C in Farads

Fig. VII-8 Constant k and m-Derived Filter Sections

derived if additional impedances are added to the shunt arm. Schematic diagrams of series derived m-type low pass filter sections, the expressions used to obtain their component values from the basic data, their basic formulas, the expressions used to obtain their component values from k-values and their curve of attenuation plotted as a function of frequency are shown in Figure VII-8. Constant k values are designed by the subscript (k) in the expressions given in Figure VII-8.

Sharpness of cut-off in the m-derived filter section is a function of m. To obtain sharp cut-off, the filter section should have a frequency of infinite attenuation, f_∞, close to the cut-off frequency, f_c. The expression

$$m = \sqrt{1 - \frac{f_c}{f_\infty}^2} \tag{3}$$

$$f_\infty = \sqrt{\frac{f_c}{1-m^2}} \tag{4}$$

which are valid for a low pass m-derived filter shows that as the ratio, f_c/f_∞ approaches unity the value of m approaches zero, or as m approaches unity the values f_∞ and f_c become more nearly equal.

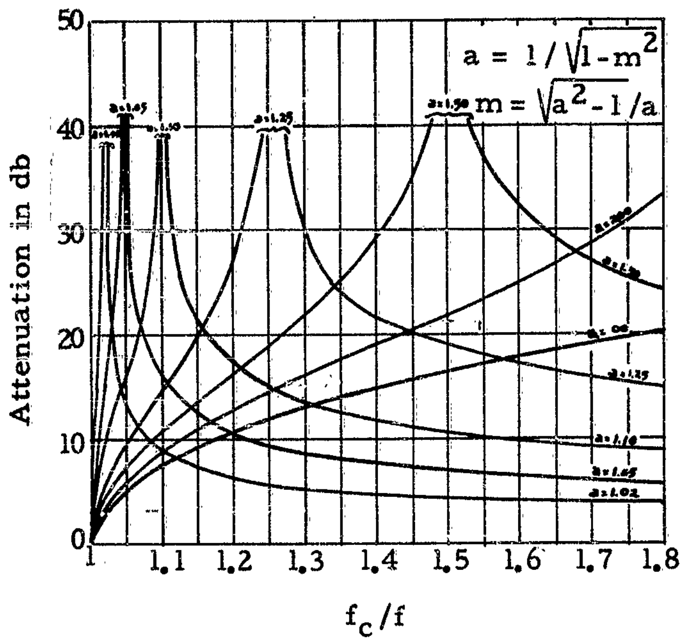


Fig. VII-9 Attenuation of m-derived Low Pass Filters

Figure VII-9 illustrates the variation of attenuation with the ratio of the cut-off frequency f_c to the interference frequency, f. The attenuation offered by m-derived filter sections for all values of m less than unity becomes infinitely high at some finite frequency, f_∞, and then decreases and approaches zero at higher frequencies. Although the sharpness of cut-off is more pronounced for the lower values of m, it is accompanied by correspondingly lower values of attenuation which approach zero

at frequencies beyond the frequency of infinite attenuation. To compensate for this undesirable characteristic a constant-k section, m = 1, is often used in conjunction with m-derived sections. Its sharpness of cut-off is not as abrupt as those of the m-derived sections, but its attenuation increases constantly and approaches infinity as the frequency increases.

In order to join two or more filter sections for the purpose of obtaining higher attenuation throughout the attenuating band, their cut-off frequencies as well as their image impedances must be identical, but their frequencies of infinite attenuation may be unequal. m-derived T sections, each with a different value of m, make this possible because the image impedances of a T section at any frequency is the same regardless of the value of m as shown in Figure VII-10. Therefore, an impedance match is obtained between the T sections and reflections will not, of course, take place. However, it is difficult to terminate these sections properly due to the variation of image impedance with frequency, but by the use of terminating half pi-sections it is possible to keep the image impedance of the filter constant at all frequencies up to approximately 90 percent of cut-off if the value of m selected is 0.6 (a = 1.25) as illustrated in Figure VII-10. The proper termination can be accomplished by designing each section as a T network and then rearranging to form a pi-section as shown in Figure VII-11. Although the entire filter looks like 3 pi-sections, the sections between the dotted lines have a T configuration. It is thus possible to alter the attenuating characteristics of a filter without varying its image impedances which are equal to the terminating impedance of the half pi-section.

Figures VII-12 through VII-15 give in tabular form values of inductance and capacitance for m-derived low pass filters having m values of 0.1, 0.2, 0.4, and 0.6, respectively. The image impedance, R, of each filter is 50 ohms. If it is necessary to employ a value of R other than the value for which the filter components are listed, the component values given in the table must first be found. The listed inductance value or values must be multiplied by R/50, while the listed capacitance value or values must be divided by R/50 in order to obtain the desired values if the filter is to be designed at some other nominal impedance.

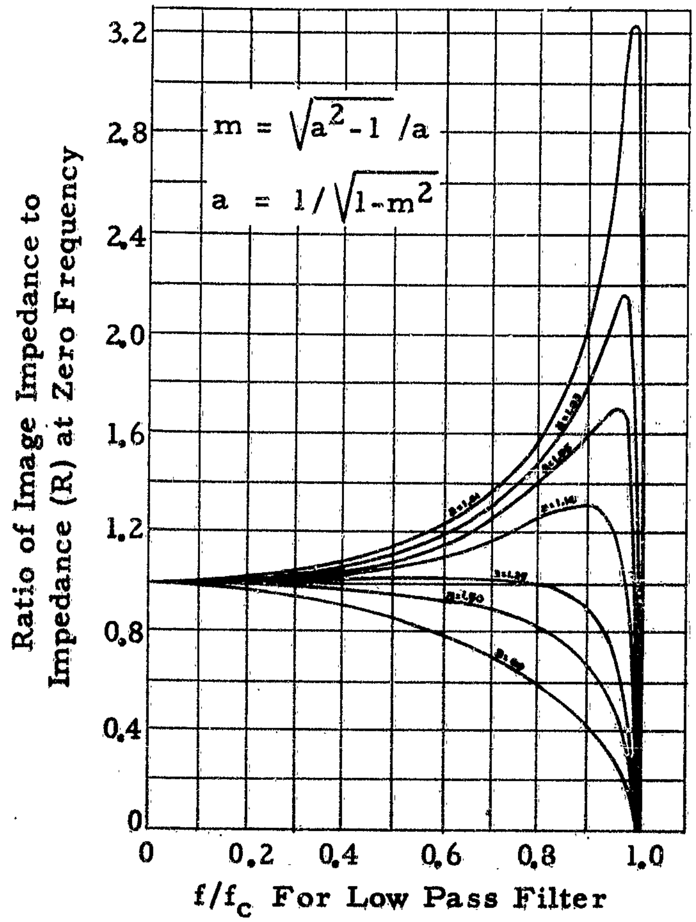


Fig. VII-10 Variation of Mid-Series and Mid-Shunt Image Impedance for Low Pass m-Derived Filter Sections

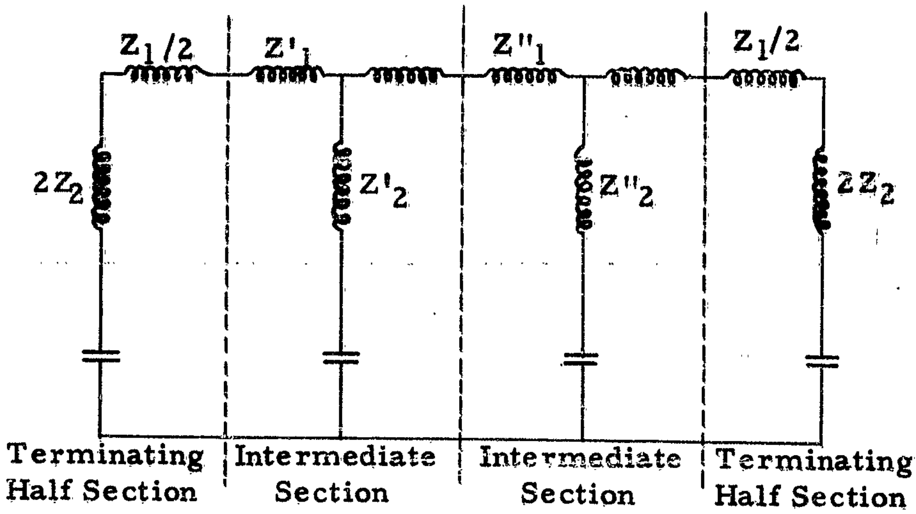


Fig. VII-11 Development of a Filter To Have a Desirable Impedance Variation

Cut-off Freq.	Series Derived			Shunt Derived		
f_c (cps)	L_1(mh)	L_2(mh)	C_2(μf)	L_1(mh)	C_1(μf)	C_2(μf)
30	5.31x10	1.31×10^{3}	2.12x10	5.31x10	5.24×10^{2}	2.12x10
100	1.59x10	3.93×10^{2}	6.37	1.59x10	1.57×10^{2}	6.37
150	1.06x10	2.62×10^{2}	4.24	1.06x10	1.05×10^{2}	4.24
200	7.96	1.97×10^{2}	3.18	7.96	7.87x10	3.18
250	6.37	1.58×10^{2}	2.55	6.37	6.30x10	2.55
300	5.31	1.31×10^{2}	2.12	5.31	5.24x10	2.12
350	4.55	1.13×10^{2}	1.82	4.55	4.50x10	1.82
400	3.98	9.80x10	1.59	3.98	3.93x10	1.59
450	3.54	8.76x10	1.41	3.54	3.49x10	1.41
500	3.18	7.87x10	1.27	3.18	3.14x10	1.27
550	2.89	7.15x10	1.16	2.89	2.87x10	1.16
600	2.65	6.56x10	1.06	2.65	2.62x10	1.06
650	2.45	6.07x10	9.79×10^{-1}	2.45	2.42x10	9.79×10^{-1}
700	2.27	5.62x10	9.09×10^{-1}	2.27	2.25x10	9.09×10^{-1}
750	2.12	5.25x10	8.49×10^{-1}	2.12	2.10x10	8.49×10^{-1}
800	1.99	4.93x10	7.96×10^{-1}	1.99	1.97x10	7.96×10^{-1}
850	1.87	4.63x10	7.49×10^{-1}	1.87	1.85x10	7.49×10^{-1}
900	1.77	4.38x10	7.07×10^{-1}	1.77	1.75x10	7.07×10^{-1}
950	1.68	4.16x10	6.70×10^{-1}	1.68	1.66x10	6.70×10^{-1}
1×10^{3}	1.59	3.93x10	6.37×10^{-1}	1.59	1.57x10	6.37×10^{-1}
3×10^{3}	5.31×10^{-1}	1.31x10	2.12×10^{-1}	5.31×10^{-1}	5.24	2.12×10^{-1}
10×10^{3}	1.59×10^{-1}	3.93	6.37×10^{-2}	1.59×10^{-1}	1.57	6.37×10^{-2}
30×10^{3}	5.31×10^{-2}	1.31	2.12×10^{-2}	5.31×10^{-2}	5.25×10^{-1}	2.12×10^{-2}
100×10^{3}	1.59×10^{-2}	3.93×10^{-1}	6.37×10^{-3}	1.59×10^{-2}	1.57×10^{-1}	6.37×10^{-3}
300×10^{3}	5.31×10^{-3}	1.31×10^{-1}	2.12×10^{-3}	5.31×10^{-3}	5.24×10^{-2}	2.12×10^{-3}
1×10^{6}	1.59×10^{-3}	3.93×10^{-2}	6.37×10^{-4}	1.59×10^{-3}	1.57×10^{-2}	6.37×10^{-4}
3×10^{6}	5.31×10^{-4}	1.31×10^{-2}	2.12×10^{-2}	5.31×10^{-4}	5.24×10^{-3}	2.12×10^{-4}
10×10^{6}	1.59×10^{-4}	3.93×10^{-3}	6.37×10^{-5}	1.59×10^{-4}	1.57×10^{-3}	6.37×10^{-5}
30×10^{6}	5.31×10^{-5}	1.31×10^{-3}	2.12×10^{-5}	5.31×10^{-5}	5.25×10^{-4}	2.12×10^{-5}

Fig. VII-12 m-Derived Low Pass Filter (R=50Ω, m=0.1)

Cut-off Freq.	Series Derived			Shunt Derived		
f_c (cps)	L_1 (mh)	L_2 (mh)	C_2 (μf)	L_1 (mh)	C_1 (μf)	C_2 (μf)
30	1.06×10^{2}	6.37×10^{2}	4.24×10	1.06×10^{2}	2.54×10^{2}	4.14×10
100	3.18×10	1.91×10^{2}	1.27×10	3.18×10	7.64×10	1.27×10
150	2.12×10	1.27×10^{2}	8.48	2.12×10	5.09×10	8.48
200	1.59×10	9.55×10	6.36	1.59×10	3.82×10	6.36
250	1.27×10	7.64×10	5.10	1.27×10	3.06×10	5.10
300	1.06×10	6.37×10	4.24	1.06×10	2.54×10	4.24
350	9.10	5.46×10	3.64	9.10	2.18×10	3.64
400	7.96	4.78×10	3.18	7.96	1.91×10	3.18
450	7.08	4.25×10	2.82	7.08	1.69×10	2.82
500	6.36	3.83×10	2.52	6.36	1.52×10	2.52
550	5.78	3.45×10	2.32	5.78	1.39×10	2.82
600	5.30	3.18×10	2.12	5.30	1.27×10	2.12
650	4.90	2.94×10	1.96	4.90	1.17×10	1.96
700	4.54	2.72×10	1.82	4.54	1.09×10	1.82
750	4.24	2.57×10	1.70	4.24	1.02×10	1.70
800	3.98	2.39×10	1.59	3.98	9.54	1.59
850	3.74	2.24×10	1.50	3.74	9.00	1.50
900	3.54	2.13×10	1.41	3.54	8.49	1.41
950	3.36	2.02×10	1.34	3.36	8.04	1.34
1×10^{3}	3.18	1.91×10	1.27	3.18	7.65	1.27
3×10^{3}	1.06	6.37	4.24×10^{-1}	1.06	2.54	4.24×10^{-1}
10×10^{3}	3.18×10^{-1}	1.91	1.27×10^{-1}	3.18×10^{-1}	7.64×10^{-1}	1.27×10^{-1}
30×10^{3}	1.06×10^{-1}	6.37×10^{-1}	4.24×10^{-2}	1.06×10^{-1}	2.54×10^{-1}	4.24×10^{-2}
100×10^{3}	3.18×10^{-2}	1.91×10^{-1}	1.37×10^{-2}	3.18×10^{-2}	7.64×10^{-2}	1.37×10^{-2}
300×10^{3}	1.06×10^{-2}	6.37×10^{-2}	4.24×10^{-3}	1.06×10^{-2}	2.54×10^{-2}	4.24×10^{-3}
1×10^{6}	3.18×10^{-3}	1.91×10^{-2}	1.27×10^{-3}	3.18×10^{-3}	7.64×10^{-3}	1.27×10^{-3}
3×10^{6}	1.06×10^{-3}	6.37×10^{-3}	4.24×10^{-4}	1.06×10^{-3}	2.54×10^{-3}	4.24×10^{-4}
10×10^{6}	3.18×10^{-4}	1.91×10^{-3}	1.27×10^{-4}	3.18×10^{-4}	7.64×10^{-4}	1.27×10^{-4}
30×10^{6}	1.06×10^{-4}	6.37×10^{-4}	4.24×10^{-5}	1.06×10^{-4}	2.54×10^{-4}	4.24×10^{-5}

Fig. VII-13 m-Derived Low Pass Filter (R=50Ω, m=0.2)

Cut-off Freq.	Series Derived			Shunt Derived		
f_c (cps)	L_1(mh)	L_2(mh)	C_2(μf)	L_1(mh)	C_1(μf)	C_2 (μf)
30	2.12×10^{2}	2.78×10^{2}	8.48×10	2.12×10^{2}	1.11×10^{2}	8.48×10
100	6.36×10	8.35×10	2.55×10	6.36×10	3.34×10	2.55×10
150	4.24×10	5.56×10	1.70×10	4.24×10	2.22×10	1.70×10
200	3.18×10	4.17×10	1.27×10	3.18×10	1.67×10	1.27×10
250	2.55×10	3.34×10	1.02×10	2.55×10	1.34×10	1.02×10
300	2.12×10	2.78×10	8.48	2.12×10	1.11×10	8.48
350	1.82×10	2.39×10	7.28	1.82×10	9.55	7.28
400	1.59×10	2.09×10	6.36	1.59×10	8.35	6.36
450	1.42×10	1.86×10	5.64	1.42×10	7.40	5.64
500	1.27×10	1.67×10	5.08	1.27×10	6.67	5.08
550	1.16×10	1.52×10	4.64	1.16×10	6.09	4.64
600	1.06×10	1.39×10	4.24	1.06×10	5.56	4.24
650	9.80	1.29×10	3.92	9.80	5.14	3.92
700	9.08	1.19×10	3.64	9.08	4.77	3.64
750	8.48	1.09×10	3.40	8.48	4.45	3.40
800	7.96	1.05×10	3.18	7.96	4.17	3.18
850	7.48	9.82	3.00	7.48	3.92	3.00
900	7.08	9.29	2.83	7.08	3.71	2.83
950	6.72	8.82	2.68	6.72	3.52	2.68
1×10^{3}	6.36	8.35	2.55	6.36	3.34	2.55
3×10^{3}	2.12	2.78	8.48×10^{-1}	2.12	1.11	8.48×10^{-1}
10×10^{3}	6.36×10^{-1}	8.35×10^{-1}	2.55×10^{-1}	6.36×10^{-1}	3.34×10^{-1}	2.55×10^{-1}
30×10^{3}	2.12×10^{-1}	2.78×10^{-1}	8.48×10^{-2}	2.12×10^{-1}	1.11×10^{-1}	8.48×10^{-2}
100×10^{3}	6.36×10^{-2}	8.35×10^{-2}	2.55×10^{-2}	6.36×10^{-2}	3.34×10^{-2}	2.55×10^{-2}
300×10^{3}	2.12×10^{-2}	2.78×10^{-2}	8.48×10^{-3}	2.12×10^{-2}	1.11×10^{-2}	8.48×10^{-3}
1×10^{6}	6.36×10^{-3}	8.35×10^{-3}	2.55×10^{-3}	6.36×10^{-3}	3.34×10^{-3}	2.55×10^{-3}
3×10^{6}	2.12×10^{-3}	2.78×10^{-3}	8.48×10^{-4}	2.12×10^{-3}	1.11×10^{-3}	8.48×10^{-4}
10×10^{6}	6.36×10^{-4}	8.35×10^{-4}	2.55×10^{-4}	6.36×10^{-4}	3.34×10^{-4}	2.55×10^{-4}
30×10^{6}	2.12×10^{-4}	2.78×10^{-4}	8.48×10^{-5}	2.12×10^{-4}	1.11×10^{-4}	8.48×10^{-5}

Fig. VII-14 m-Derived Low Pass Filter ($R=50\Omega$, $m=0.4$)

Cut-off Freq.	Series Derived			Shunt Derived		
f_c (cps)	L_1(mh)	L_2(mh)	C_2 (μf)	L_1(mh)	C_1(μf)	C_2(μf)
30	3.19×10^{2}	1.41×10^{2}	1.27×10^{2}	3.19×10^{2}	5.64×10^{2}	1.27×10^{2}
100	9.54×10	4.23×10	3.82×10	9.54×10	1.69×10	3.82×10
150	6.36×10	2.82×10	2.54×10	6.36×10	1.13×10	2.54×10
200	4.78×10	2.12×10	1.91×10	4.78×10	8.45	1.91×10
250	3.82×10	1.69×10	1.53×10	3.82×10	6.78	1.53×10
300	3.19×10	1.41×10	1.27×10	3.19×10	5.64	1.27×10
350	2.73×10	1.21×10	1.09×10	2.73×10	4.84	1.09×10
400	2.39×10	1.06×10	9.54	2.39×10	4.22	9.54
450	2.12×10	9.41	8.46	2.12×10	3.75	8.46
500	1.91×10	8.45	7.62	1.91×10	3.37	7.62
550	1.73×10	7.68	6.96	1.73×10	3.08	6.96
600	1.59×10	7.04	6.36	1.59×10	2.81	6.36
650	1.47×10	6.51	5.87	1.47×10	2.60	5.87
700	1.36×10	6.03	5.45	1.36×10	2.41	5.45
750	1.27×10	5.63	5.09	1.27×10	2.25	5.09
800	1.19×10	5.29	4.78	1.19×10	2.12	4.78
850	1.12×10	4.97	4.49	1.12×10	1.99	4.49
900	1.06×10	4.70	4.24	1.06×10	1.88	4.24
950	1.01×10	4.46	4.02	1.01×10	1.78	4.02
1×10^{3}	9.54	4.23	3.82	9.54	1.69	3.82
3×10^{3}	3.19	1.41	1.27	3.19	5.64×10^{-1}	1.27
10×10^{3}	9.54×10^{-1}	4.23×10^{-1}	3.82×10^{-1}	9.54×10^{-1}	1.69×10^{-1}	3.82×10^{-1}
30×10^{3}	3.19×10^{-1}	1.41×10^{-1}	1.27×10^{-1}	3.19×10^{-1}	5.64×10^{-2}	1.27×10^{-1}
100×10^{3}	9.54×10^{-2}	4.23×10^{-2}	3.82×10^{-2}	9.54×10^{-2}	1.69×10^{-2}	3.82×10^{-2}
300×10^{3}	3.19×10^{-2}	1.41×10^{-2}	1.27×10^{-2}	3.19×10^{-2}	5.64×10^{-3}	1.27×10^{-2}
1×10^{6}	9.54×10^{-3}	4.23×10^{-3}	3.82×10^{-3}	9.54×10^{-3}	1.69×10^{-3}	3.82×10^{-3}
3×10^{6}	3.19×10^{-3}	1.41×10^{-3}	1.27×10^{-3}	3.19×10^{-3}	5.64×10^{-4}	1.27×10^{-3}
10×10^{6}	9.54×10^{-4}	4.23×10^{-4}	3.82×10^{-4}	9.54×10^{-4}	1.69×10^{-4}	3.82×10^{-4}
30×10^{6}	3.19×10^{-4}	1.41×10^{-4}	1.27×10^{-4}	3.19×10^{-4}	5.64×10^{-5}	1.27×10^{-4}

Fig. VII-15 m-Derived Low Pass Filter (R=50Ω, m=0.6)

APPENDIX VIII

SHIELDED ROOM - CONSTRUCTION AND USE

The shielded room is designed and constructed to provide a working area free of spurious electromagnetic energy where such energy would interfere with the proper operation of equipment or measurements being made.

The shielding of one area from another is normally obtained by placing a physical barrier between the two regions. This shielding can be regarded as resulting from reflection and attenuation of the incident energy. Good reflection is dependent primarily upon high conductivity, and good attenuation upon a large conductivity-permeability product. A more detailed discussion of these quantities is given in Appendix XVI.

The majority of shielded rooms have been built of wire mesh; however, in order to obtain effective shielding at very high frequencies, sheet metal is used. Many high-quality rooms are double-walled and of sheet copper. Copper, however, is expensive and difficult to obtain in times of emergency. Certain iron and steel materials are more readily obtainable, cheaper, and theoretically able to satisfy the electrical requirements with a reasonable wall thickness.

If mesh walls are used, shielding is due primarily to reflection loss. Copper screening with at least sixty strands per wave length at the highest operating frequency is the best to use. For optimum results it is well to make sure that the various overlapping wires of the mesh make good electrical contact with one another; they may, for instance, be soldered together. If this is not done, corrosion may occur at the overlap point and cause each corroded joint to act as a noise generator.

Shielded rooms may have either a single wall or double walls. Rooms that have an outer wall not in metallic contact with the inner wall are said to be of the "doubly shielded type". When the two walls are electrically connected, the room is said to be of the "cell type".

Cell-type mesh wall shielding rooms are commercially available and are reported to give 100 db shielding from 0.15 to 10,000 megacycles. Cell-type rooms have the advantage of portability since they are not ordinarily soldered together.

On the other hand, if solid metal walls are used, where a large part of the shielding is due to absorption loss, iron walls are best. For example, 24 gage galvanized iron sheet 0.635 millimeters thick gives, on the basis of an assumed relative permeability of 1000, and absorption loss of 133 db at 13 kilocycles and greater loss at higher frequencies.

SHIELDING VALUE OF VARIOUS SHIELDS

(a) Single Solid Metal Shields

For a single solid metal shield, the total loss in decibels, assuming plane

waves, is given by equation (21), Appendix XVI:

$$\text{Total shielding loss} = 3.34\sqrt{f_m \mu_r \sigma_r} \quad S + 108.2 + 10 \log \frac{\sigma r}{f_m \mu_r}$$

where f_m is the frequency in megacycles per second, μ_r the relative magnetic permeability of the shielding material ($\mu_r = 1$ for all non-magnetic materials), σ_r the relative conductivity of the shielding material ($\sigma_r = 1$ for copper), and S the thickness of the shield in mils. This equation is plotted for copper and some magnetic material in Figures XVI-3 and XVI-4.

(b) Single Wire Mesh Shields

Shields may be of wire mesh provided that the individual strands are joined to one another at their points of intersection, and that the size of the mesh openings is extremely small compared to a wavelength. Under these conditions the mesh acts as a surface with an impedance approximately the same as that of the metal composing the strands. The principle shielding action of such a mesh is due to the reflection at the first surface and the shielding loss in decibels is therefore given by the expression:

$$\text{Shielding Loss} = 10 \log \frac{(Z_m + Z_{air})^2}{4 Z_m Z_{air}} \tag{1}$$

Where Z_{air} is the intrinsic impedance of air for the type of wave considered, and Z_m is the impedance of the mesh.

The mesh impedance Z_m is a quantity that is very difficult to determine analytically since it depends in a complicated way on the size of the wire, the size of the openings between wires, and the impedance of the wire itself. It is found that the mesh with 50% open area, and on the order of sixty or more strands per wavelength, has an impedance very nearly the same as that of the mesh material itself. Experimental confirmation of this result has been tabulated below in Figure VIII-1.

Method of obtaining data	Shielding-db	Material	λ(cm)	Strands per wavelength
Measured	23	copper screen	9.1	17.8
Measured	33	copper screen	9.1	48.5
Calculated	39	copper sheet	9.1	----
	(Reflection one surface only)			

Fig. VIII-1 Shielding of Copper Sheet and Mesh

(c) Double Shields

Shielded rooms that have an outer wall that is not in metallic contact with the inner wall have a greater shielding efficiency than single-wall rooms or rooms that have inner and outer shields connected in many places. The shielding of the electrically isolated double-wall rooms depends not only on the wall material but also on the spacing between the walls. It turns out that a given thickness of metal divided into two walls gives more shielding than the same metal in a single wall.

If the inner and outer walls are conductively connected at only one point, for example where the power line enters the room, the walls are still effectively decoupled since there is no "return path" for the conduction current. If connection is made between the inner and outer wall at two or more points, then the same current can flow in the inner and outer walls and the shielding improvement due to the separation of the walls is lessened, but the shielding effectiveness is greater than that of a single wall with the same amount of metal.

(1) Solid Metal Walls

The shielding action of the double-wall room with one or zero points conductive coupling between the inner and outer walls may be analyzed as follows: The absorption loss of each wall is as given for single solid metal walls. The reflection loss in decibels, assuming that the absorption loss in each wall is greater than 10 db, is given in general as:

$$\text{Total Reflection Loss} = 20 \log \frac{(Z_m + Z_{air})^2}{4Z_m Z_{air}}$$

$$+ 20 \log \frac{\dfrac{Z_m + Z_{air}}{4Z_m Z_{air}} \cdot \dfrac{\dfrac{Z_m + jZ_{air} \tan 2 L/\lambda}{Z_{air} + jZ_m \tan 2 L/\lambda}}{\dfrac{Z_m + jZ_{air} \tan 2 L/\lambda}{Z_{air} + jZ_m \tan 2 L/\lambda}}}{} \qquad (2)$$

where L is the distance between shields in meters and λ is the wavelength.

When the spacing of the shields is very close or when $L/\lambda = N/2$, where $N = 1, 2, 3, \ldots$ the second term of the equation approaches zero. These are conditions of minimum reflection. When the spacing is an odd number of quarter wavelengths, that is $L/\lambda = (2N + 1)/4$, where $N = 0, 1, 2,$ etc., then $\tan (2\pi L/\lambda) \to \infty$ and the general equation becomes

$$\text{Total Reflection Loss} = 20 \log \frac{(Z_m + Z_{air})^2}{4Z_m Z_{air}} + 40 \log \frac{Z_m + \dfrac{Z_{air}}{Z_m}^2}{2\, Z_{air}} \qquad (3)$$

These are the conditions of maximum reflection loss.

(2) Mesh Walls

For the case of two mesh screens electrically isolated, or connected at one point, the predominant part of the shielding is a result of reflection loss which will depend upon the spacing between mesh screens, as measured in

wavelengths. The reflection loss is the same as in case (a) with Z_m now standing for the impedance of the mesh, provided only that the mesh impedance is much smaller than the impedance of air.

(d) Cell-Type Shields

(1) Mesh Walls

Cell-type shielded rooms have been in common use for seven or eight years, and give quite satisfactory results for most applications. In this type of construction the inner and outer walls are made of a mesh and joined together at many points. It is impossible to give a simple quantitative picture of the behavior of such a shield. However, there are several general things which may be said qualitatively about its method of operation. At very low frequencies, where the dimensions of the room are small compared to a free space wavelength, the inner and outer shields are effectively in parallel and so one should expect to find the reflection loss given by

$$\text{Reflection Loss} = 10 \log \frac{\left(Z_{air} + \frac{Z_m}{2}^2 \right)}{2Z_{air}\, Z_m} \tag{4}$$

Thus, at low frequencies one expects the shielding to be at least 3 db better than a single shield. At very high frequencies, where the dimensions of the room are very large compared to the wavelength, the multiple connection points should have little effect on the shielding of the room and so Equation (2) and Equation (3) should describe the shielding properties. Therefore, the cell-type shielded room with double mesh walls can always be expected to give better shielding than a room with single mesh walls, but not as good shielding as the double shielded room with mesh walls.

(2) Solid Metal Walls

Solid-wall rooms can be built using a cell-type construction. As with the cell-type rooms with mesh walls, it is impossible to give a completely quantitative picture of the shielding action, but a qualitative picture of the behavior can be given. At frequencies near 15 kilocycles, where the absorption loss is smallest, there is some conductive coupling between the inner and outer walls. However, as can be seen by reference to Figures XVI-3 and XVI-4, the absorption loss for any reasonable thickness of material is great enough so that the inner and outer shields are fairly loosely coupled. Therefore, to a first approximation, the conductive coupling between shields can be neglected and shielding action can be considered to be almost the same as that of a double-wall room. At very high frequencies the approximation of no conductive coupling between inner and outer shields is a very good one and the shielding is more nearly identical with that of the double-wall room.

CONSTRUCTION DETAILS

(a) Framework

The structural framework on which the shield material is placed is commonly

constructed of a good insulator such as dry, varnished spruce or redwood. Redwood has been used as the insulating material in high voltage impulse generators and has given very satisfactory results. A good insulator is mandatory for the framework of a double-wall room to prevent the flow of circulating conduction current. During construction it is well to take resistance readings with a megohmmeter between inner and outer walls to make sure that no extraneous conduction paths occur. For a single-wall room or cell type room either an insulator or a conductor may be used for the framework.

(b) Wall Fastening

The joining of one section of wall material to another must be carefully done in order to realize optimum shielding. In the case of solid metal walls, lap joints are to be preferred with at least a half-inch overlap. Any method of joining the laps together in a continuous fashion, such as soft soldering or roll welding is satisfactory; spot welding does not give good results. Screen walls can be lapped and soldered, or they can be butt-jointed as shown in Figure VIII-2. Unsoldered butt joints have been found to give very good results and are therefore to be recommended for portable installations. Soldered joints, because of their greater mechanical strength, are better for permanent installations.

The shielding walls may be fastened to the structural framework in a variety of ways. Some methods of fastening screen walls are shown in Figure VIII-2 for butt-jointed sections. Where lap joints are used on screening, U nails may be employed to connect the screening to wooden structural members; soldering may be used if the structural members are metal.

Solid metal walls may be fastened to a wooden framework by means of large-head roofing nails, provided each nail head is soldered to the wall. If a metal framework is used, spot welding may be employed.

(c) Doors

The doors for the shielded room should be constructed on the inside and outside of the same material as the inside and outside walls of the room. The door must not make electrical contact between the inner and outer walls if double-wall construction is to be used. This restriction, however, does not apply to cell-type construction. Ordinary 10-mil thick phosphor bronze weather stripping should be placed on the inside edges of the door and jamb so that door and jamb make good electrical contact when the door is closed. If the door and jamb are beveled it will help in making the door electrically tight.

The door may be fastened securely by means of several latches which wedge the door closed as they are rotated. It is good practice to use several latches in order that the door does not spring. An alternative approach to using multiple latches is to use a loop of thin-wall rubber tubing set in the door jamb completely around the perimeter of the door. This tubing should be filled with a relatively non-compressible fluid such as water. The door handle can be fitted with a cam that forces a plate against the inside edge of the rubber tubing, compressing it as the latch is rotated. This device wedges the weather stripping together very tightly. Satisfactory doors have been built where the rubber tubing is expanded by means of compressed air

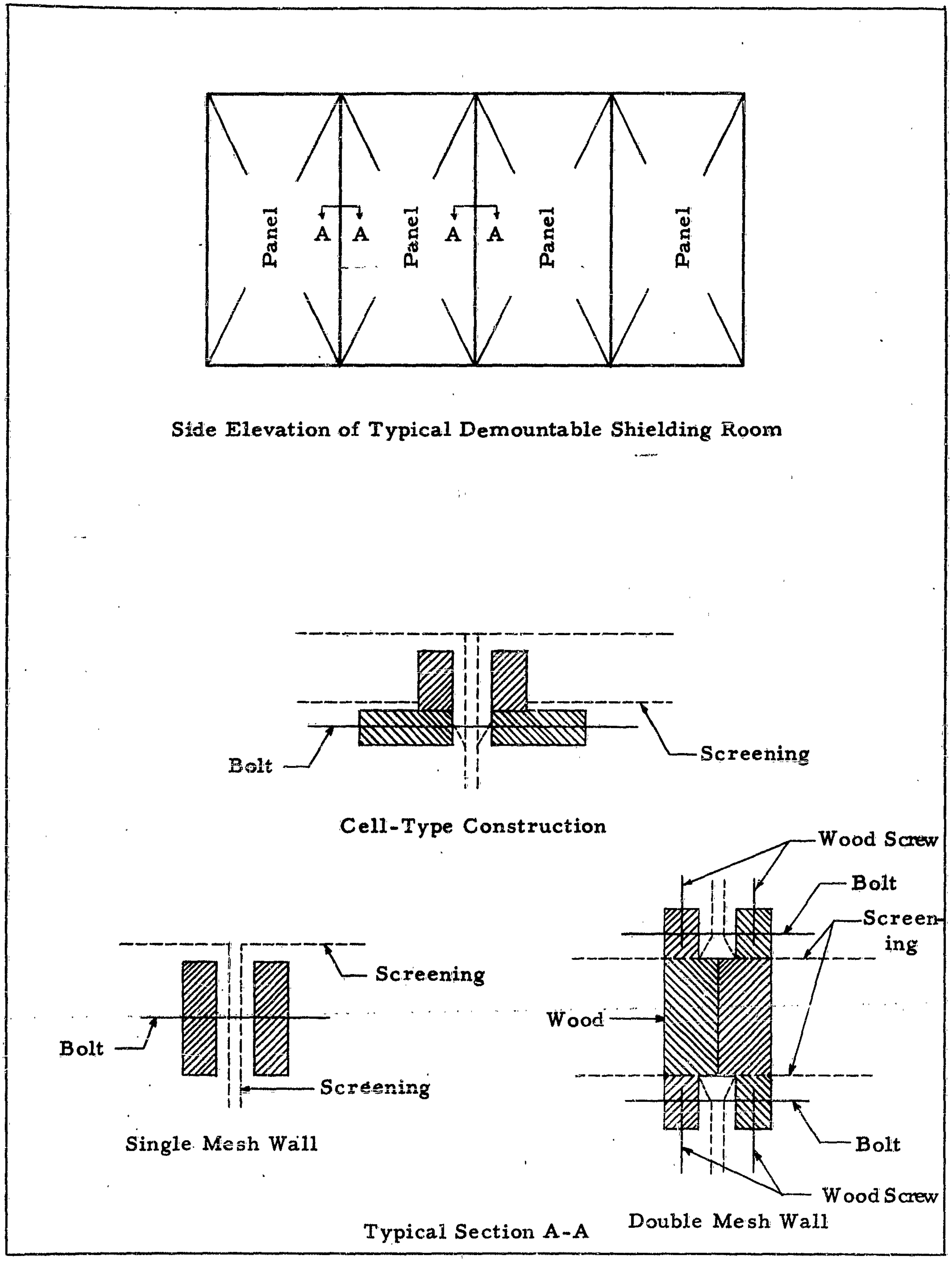


Fig. VIII-2 Methods of Joining Mesh Wall Panels for Demountable Construction

Figure VIII-3 shows a typical arrangement of the components of the door and door jamb for a room with an inner and outer shield of sheet iron. A door similar to that pictured in Figure VIII-3 can be used with a single-wall room if the outer half of the shielding is removed.

(d) Ventilation

The ventilation of screen rooms is no problem since the air can circulate freely through the mesh. Air can be introduced into solid-wall rooms by means of a honeycomb arrangement that acts as a wave guide below cut-off to the radio waves. (See Paragraph 3.1.2.2 for the design of wave guides.) One easy way to obtain the honeycomb is to use an automobile radiator. If this is not available the honeycomb can be built with an "egg-crate" type construction. Each cell must be joined electrically to its neighbor or to the frame by soldering or an equivalent process. The frame, in turn, must be connected to the metal wall by means of an L channel soldered to the frame on one side and to the wall on the other. The honeycomb can be constructed from any convenient material. It is obvious that such filters must be placed in both shields in rooms built with double solid metal walls. These filters, of course, must be electrically isolated.

(e) Services

Various services may be brought into the room without affecting its shielding properties. Services such as water or gas which are conveyed by means of metal pipes can be brought through the walls of a double-shielded room near the power lines without causing difficulty due to multiple connections. In single-wall rooms or cell-type rooms the point of entry of the services is immaterial. In all type rooms, however, it is necessary to bond the pipes firmly to the walls.

Lighting can be obtained for screen rooms by means of lights located outside the room. Lighting for solid metal wall rooms should be obtained by means of incandescent lamps run from the room's power supply. Flourescent lights are not recommended because they have a high interference power output throughout the radio-frequency spectrum.

(f) Power Line Filters

It is necessary to have power available inside a shielded room to operate test equipment. One method of obtaining power inside the room is by means of batteries. However, this is costly and an inconvenient source of power. Another method is to drive an alternator inside the room by means of a shaft extending through the room wall. If the drive shaft is made of metal, energy at all frequencies can come in between the shaft and its bearing because this combination acts as a coaxial line. If the drive shaft is made of a non-conductor, then the combination will act as wave guide below cut-off at low frequencies, but at high frequencies will propogate energy readily for any reasonably sized shaft. The most practical way to get power into the room is to bring it in on well-filtered power lines.

The power-line filter is the most critical component in the shielded room. Usually the over-all shielding characteristics of a room are the same as those of the filter. This reflects the fact that the filter is the most difficult component of the room to build. (See Paragraph 3.1.1.2 for design of power line filters.)

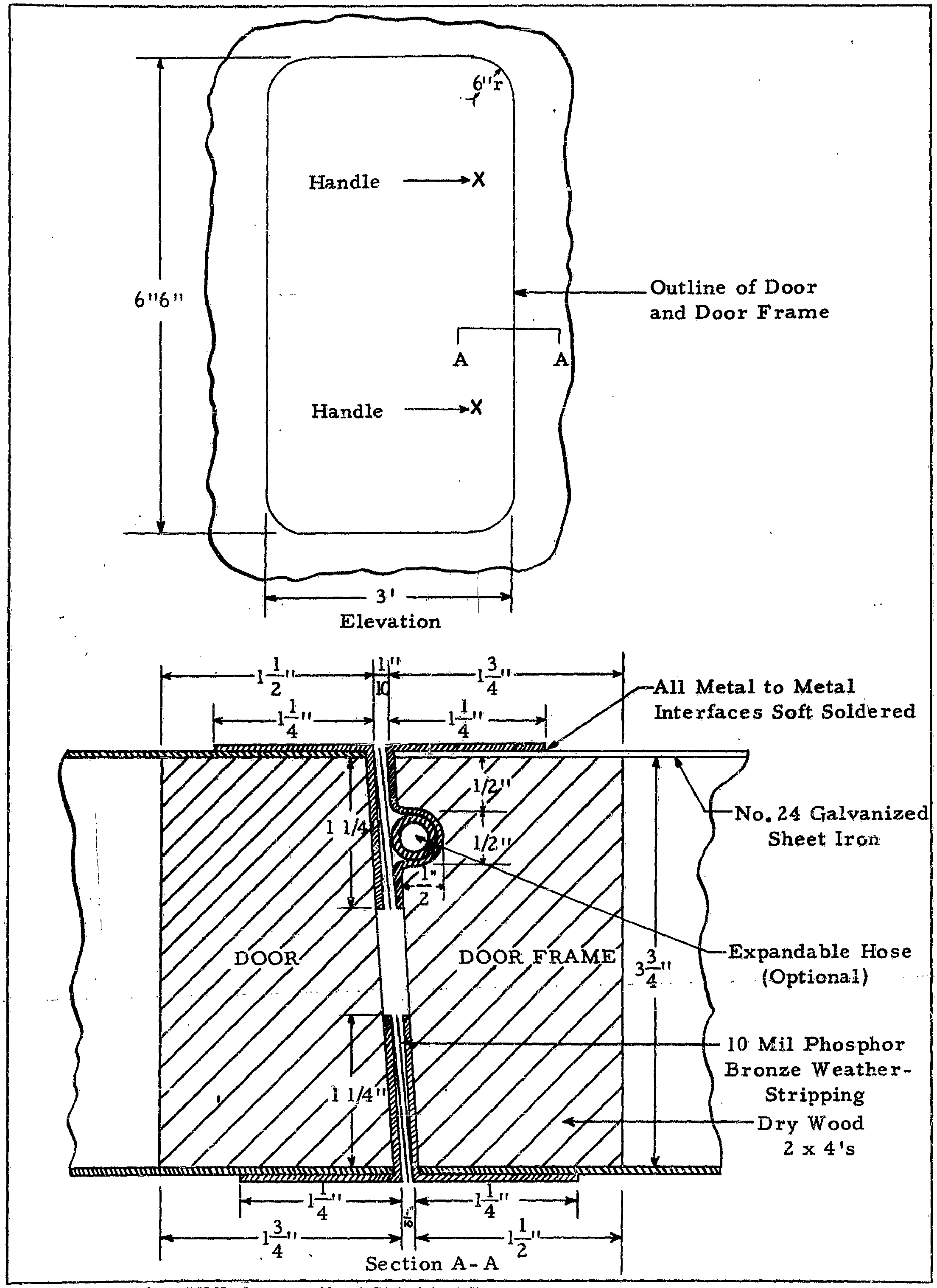


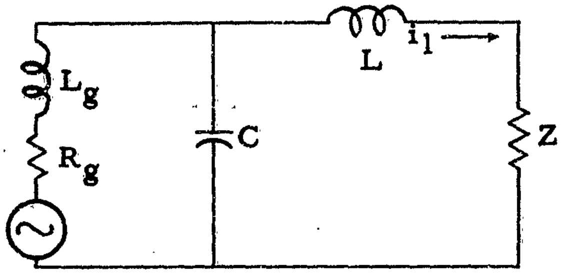
Fig. VIII-3 Detail of Shielded Room Door and Door Frame

APPENDIX IX

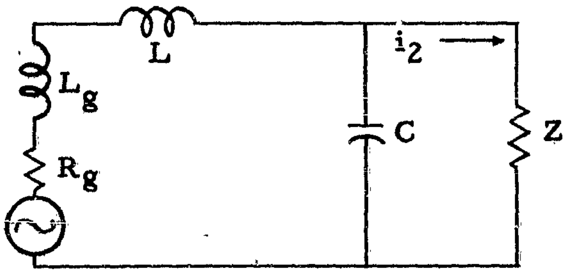
COMPARISON OF CONDENSER INPUT AND INDUCTANCE INPUT

L-TYPE NETWORKS FOR USE IN DC MOTORS

It is required to determine the proper position of a condenser to be installed in a direct current motor with a series field. Assume that the interference source acts as a voltage generator with an inductive internal impedance. Consider the following two networks and their matrices.



$$\begin{pmatrix} R_g + j\left[\omega L_g - \frac{1}{\omega C}\right] & \frac{j}{\omega C} \\ \frac{j}{\omega C} & Z + j\left[\omega L - \frac{1}{\omega C}\right] \end{pmatrix}$$



$$\begin{pmatrix} R_g + j\left[\omega(L_g + L) - \frac{1}{\omega C}\right] & \frac{j}{\omega C} \\ \frac{j}{\omega C} & Z - \frac{j}{\omega C} \end{pmatrix}$$

L is the inductance of the series field. Z is the load impedance seen by the motor, which may be resistive, inductive, or capacitive. Hence, the ratio of the currents is:

$$\frac{i_2}{i_1} = \frac{\left\{R_g + j\left[\omega L_g - \frac{1}{\omega C}\right]\right\}\left\{Z + j\left[\omega L - \frac{1}{\omega C}\right]\right\} + \left(\frac{1}{\omega C}\right)^2}{\left\{R_g + j\left[\omega(L_g + L)\frac{1}{\omega C}\right]\right\}\left\{Z - \frac{j}{\omega C}\right\} + \left(\frac{1}{\omega C}\right)^2} \qquad (1)$$

Let $Z = R + jX$. Then the ratio reduces to:

$$\frac{i_2}{i_1} = \frac{A - \omega L_g + j(B + R_g)}{A - X + j(B + R)} \qquad (2)$$

where $$A = \frac{R_g R}{\omega L} + \frac{1}{\omega C} + \frac{X}{\omega^2 LC} + \frac{L_g}{\omega LC} - \frac{L_g X}{L}$$

and $$B = \frac{R_g X}{\omega L} + \frac{L_g R}{L} - \frac{R_g + R}{\omega^2 LC}$$

Now the question is whether the ratio $|i_2/i_1|$ is larger or smaller than unity. Consider the following cases:

(a) The load is resistive; $X = 0$. Then the absolute value of the numerator is larger than that of the denominator for high frequencies, say $\omega > 10^5$, since A decreases rapidly with frequency and the imaginary parts do not differ from one another by much for normal values of R_g and R. Hence, for the frequencies of interest,

$$\left|\frac{i_2}{i_1}\right| > 1$$

(b) The load is inductive; $X = L_L > 0$. Then, for small A, the denominator increases as ωL_L and the numerator as ωL_g. L_g, the armature inductance of the motor, is normally about 1 to 5 henries. No load inductance as large as this is likely to be encountered. Hence it may be assumed that $L_L < L_g$, and again

$$\left|\frac{i_2}{i_1}\right| > 1$$

(c) The load is capacitive; $X = -(1/\omega C_L) < 0$. Then, for small A, the numerator increases as ωL_L and the denominator as $1/\omega C_L$. For normal values of C_L (10^{-11} to 10^{-7} farads) and the frequencies of interest, the numerator will again be larger than the denominator. Hence, again

$$\left|\frac{i_2}{i_1}\right| > 1$$

It is concluded that the condenser arrangement of the first circuit shown is always preferable except for those rare cases where the above assumptions do not hold.

APPENDIX X

MEASUREMENT OF INSERTION LOSS OF FILTERS

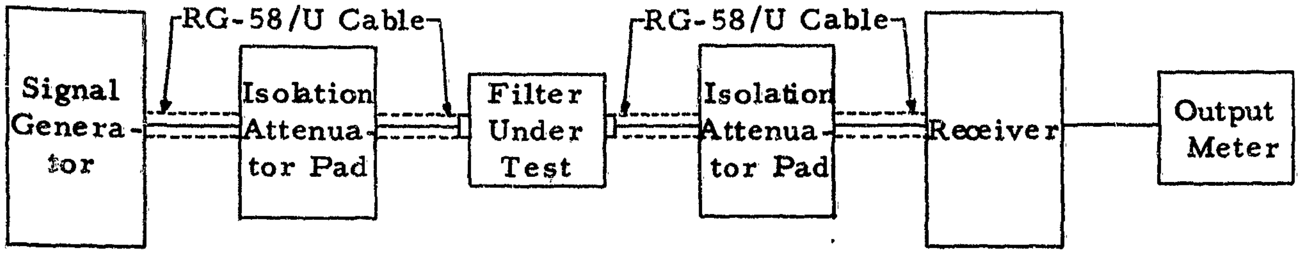
This appendix covers the most recent technique and procedure for measuring the insertion loss of filters designed to operate in the frequency range of 10 to 1000 megacycles. Insertion loss is defined as the number of decibels, or nepers, by which the current on the load side of a network has been changed by the insertion of the filter. Since the current through the load is reduced by the insertion of a shunt path, load voltage is proportionately reduced, assuming a constant load. The ratio of output current with a shunt path to the output current without a shunt path is equivalent to the ratio of voltage across the load with and without the shunt path. Therefore, insertion loss is measured more conveniently by the expression 20 log E_2/E_1. The ratio, E_2/E_1 is explained in a later paragraph.

Accurate insertion loss measurement requires that the filter under test be inserted between source and load impedances which remain constant and are known throughout the entire frequency range of measurement. Isolation pads or attenuators are required to provide sufficient isolation between the filter under test and the signal generator, as well as the signal detector on the output side. This provides a constant and specific impedance to the filter at all frequencies. A 50 ohm resistive network has been adopted for convenience, since 50 ohm cable and connectors are standard items.

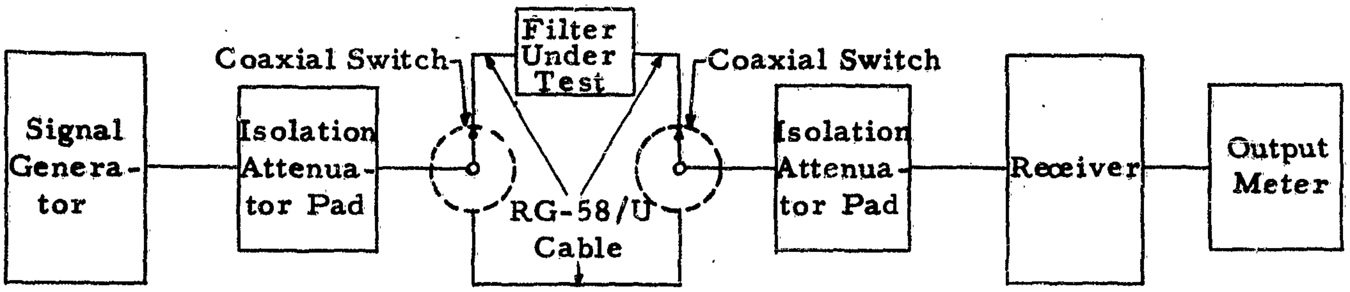
Figure X illustrates two arrangements for measuring insertion loss. The alternate method given in b is useful for obtaining measurements up to 400 megacycles. The basic circuit, as shown, consists of a calibrated RF sine wave voltage source which supplies a signal through a coaxial transmission line and attenuator to the filter under test, then to another attenuator, receiver, with the final output registered on the output meter. In the process of obtaining data for measurement, as a first step adjust the receiver gain for a convenient output indication of receiver fluctuation interference. Then apply power to the signal generator, and adjust the output indication 1 decibel above the interference level. This will give the output voltage, E_1, for the filter-out condition. After inserting the filter for measurement, first retune the receiver to resonance, and again adjust the output of the signal generator for the same receiver output level as indicated on the output meter. This output of the signal generator yields the value of E_2, and the insertion loss of the filter at the frequency of measurement will be given by the expression, as previously noted, 20 log E_2/E_1.

The accuracy of measurement will be affected by several factors. Error is commonly introduced by the variation of impedance with frequency between the attenuators and filter under test. This is probably due to a slight mismatch between attenuator impedance and line impedance, and an impedance variation will occur at the points where the line connects to the filter unit. Maximum variation would occur at the frequency for which the electrical line length is 1/4 of a wavelength or any odd multiple thereof. The attenuator may well have a frequency characteristic causing its impedance to vary with frequency, so that the impedance variation at the cable end will combine the two effects. The overall impedance variation, up to 1000

megacycles, can be held to 20 percent of the low frequency value, allowing error of 1.6 db. Up to 400 megacycles, these effects combine to produce an error less than 0.2 db.



(a) Basic Test Circuit



(b) Alternate Test Circuit

Fig. X Circuit for Measuring Insertion Loss of Filters

Some leakage from the signal generator, from poor cable, or from connectors may be picked up by direct radiation or by some unintentional coupling to the receiver antenna circuit. Properly shielded and filtered instruments are the best possible solution to this problem. In the case of undesirable coupling occurring with a given test set-up, receiver and signal generator must be separated as far as possible. To obtain good results, make sure that all radio frequency connectors are tight, all cables are well shielded, and then operate at a signal level and sensitivity which reduces any possible extraneous coupling to a minimum.

Where multiple-filter circuits are to be measured, the procedure is to measure one section leaving all others open. Then measure the same section short circuiting all others. The short circuit connections shall be as short and direct as possible. The lesser of the two measurements will be considered the insertion loss of the circuit at the test frequency.

APPENDIX XI

METHODS OF MEASURING THE EFFECTIVENESS OF SHIELDS

All practical shields consist of metallic walls, which may be solid or may consist of mesh or braid, separating two regions of space. The effectiveness of a shield is due to its ability to isolate electromagnetic phenomena in one of these two regions. This effectiveness depends not only on the shield itself, i.e., its shape, material and physical dimensions, but also on the type of electromagnetic waves used during the measurements, i.e., their impedance (ratio of electric to magnetic field intensity), frequency, and polarization. Therefore there is no absolute measure of the effectiveness of a shield. Measurements can determine only the relative shielding effectiveness under a given set of test conditions.

No standard methods are available for measuring the effectiveness of shields used to enclose completely interference-generating units such as motors, or interference-susceptible units such as receivers. Here effectiveness must be determined by actual operation. The motor is run in the way that resembles as closely as possible actual operating conditions and the region outside the shield is explored with suitable pick-up devices at all frequencies of interest. The shielding is considered effective if no signal can be detected. A receiver may be tested similarly with a strong interference source placed directly outside and precautions taken that the signal cannot enter any other way than through the shield.

The testing of cables and conduit for shielding effectiveness is particularly important in connection with the suppression of radio interference. Special standard test methods have been developed, two of which will be described in detail here.

Briefly, the first method consists of measuring the voltage drop on the outside of the shield when a specified current flows through the conductor or conductors inside the shield. This method is based on the fact that, in order for electric and magnetic fields to exist outside a region completely enclosed by the metallic shield and containing no other sources, currents and charges must be present on the surface of the shield on the side facing that region. A perfect shield would restrict all currents and charges to its inside; hence, no fields could be present in the outside region. The voltage drop along the outside of the shield, which is a measure of the integrated effect of the currents and charges on the outside, is, therefore, a measure of the effectiveness of the shield. This method is particularly suited for coaxial cables consisting of one inner conductor and a concentric outer sheath serving both as shield and as return path for the current. It is also applicable to conduit carrying more than one conductor, but a difficulty arises because it is not immediately clear what is meant by "the current inside" if there are several conductors carrying different currents, possibly in opposite directions. It is, however, the only possible method for conduit filled with a solid dielectric, in which the conductors are embedded, since the second method requires the replacement of the inner conductors by a radiating coil.

The second method consists of measuring the field strength at a point in the vicinity of the specified source both with and without the source being enclosed by

the shield to be evaluated. This may be called a "direct" method of measurement since the ratio of the two field strengths is a direct measure of the attenuation introduced by the shield. It is a measure, however, only for the particular type of field produced by the source, which may be quite different from the actual fields encountered in practice. This method is particularly suited to shielding conduit with more than one inner conductor in which the shield carries little or none of the line current. It is not directly applicable to shields carrying the return current because, when the shield is used as a return path, removal of the shield destroys the circuit, and, hence, the conditions could not be kept similar for measurements with and without the shield. Also, this method can obviously not be used when the conduit is filled with a solid dielectric.

A detailed description of the two methods follows.

Method I

1. Principle of Operation

This method utilizes the concept of surface transfer impedance, which is defined as the longitudinal voltage drop along the outside of the shield per ampere of current carried by the shield. The impedances of primary interest are small; hence, units are given in microhms. Comparative tests are referred to a standard. A tube as described in Figure XI-1 is recommended. For best results, the test specimen and the standard should have the same size. Relative leakage of zero db indicates leakage equal to that from the standard.

The lower the transfer impedance per unit length, the better the shield. Thus its reciprocal, the transfer admittance, would more appropriately represent a unit expressing shielding effectiveness.

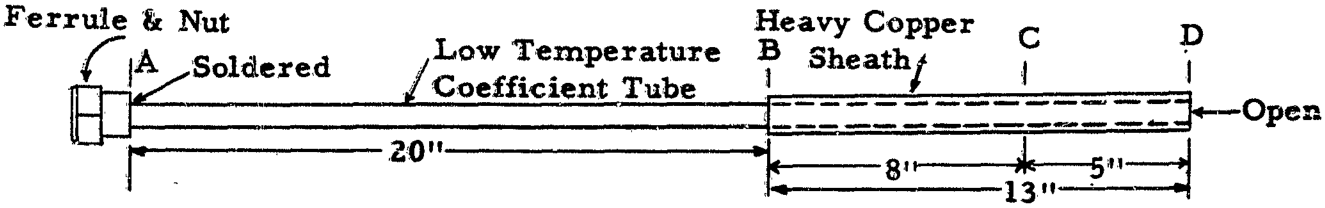
Longitudinal external voltage drops, E_x and E_s, across specimen and standard, respectively, are compared with the same currents and at the same frequency. Leakage is expressed in decibels and computed from the equation

$$\text{Leakage of specimen relative to a standard} = 20 \log_{10} \frac{E_x}{E_s} \qquad (1)$$

For shielding effectiveness, the same equation is used with the voltage ratio inverted.

A calibrated radio-frequency microvoltmeter can be used to find the external voltage drop across the specimen. However, it is preferred to evaluate this voltage by comparison with a signal of equal intensity from a standard signal generator. By means of a switch, a suitable receiving device receives the signal from specimen and generator, alternately, and the generator is adjusted until equal receiver outputs are obtained.

To measure transfer impedance directly, it is necessary to measure the current in the center lead. The basic circuit is indicated in Figure XI-2. A current indicating device, such as a shielded radio-frequency ammeter, is inserted in series with the inner conductor of the coaxial line of which the specimen forms the outer conductor. The voltage, read from the signal generator setting in the same way as



Instructions

(a) Tube made of 70-30 Cupro-nickel, a non-magnetic alloy, which has an attenuation of about 21 percent and a thermal coefficient of attenuation of about 1.2 percent of that of copper.

(b) Change in attenuation is absolutely negligible for ordinary temperature variations.

(c) A tube of this alloy is usable as a standard (to provide a given amount of attenuation) at frequencies more than 20 times as high as is a copper tube of the same wall thickness.

(d) Wall thickness should be made as uniform as possible. A 0.125-inch-thick tube is suitable for radio frequencies up to about 1 mc; a thickness of 0.050 inches is satisfactory from 0.1 to about 8 mc.

(e) A straight stiff center conductor (preferably 1/16 to 1/8-inch diameter hard copper) is held centrally by means of 4 or 5 equally-spaced thin polystyrene wafers.

(f) Connection from the center conductor to the inside of the tube is made at point "C". The sheath extension beyond the point of contact C (indicated by the section CB) permits an equal current distribution around the periphery; without this extension lack of perfectly symmetrical contact around the entire circumference would result in a non-symmetrical current distribution.

(g) Contacts for measuring leakage are applied across section AB; about 20 inches is a convenient length.

(h) A heavy copper sheath BD (at least 1/16-inch thick) is used to reduce leakage from section BD to a point where it is negligible compared with the leakage from test section AB. The sheath fits snugly over the cupro-nickel tube and is sweated on with soft solder.

(i) Section CD acts as a cut-off tube preventing leakage from the open end D. Its length should be at least five times the inside diameter of the inner tube.

(j) The standard has a spherical-faced brass ferrule soldered to end A. The assembly fits into a conically beveled seat and is secured by means of the nut shown in the figure.

Fig. XI-1 Diagram and Instructions for a Tubular Shielding Standard

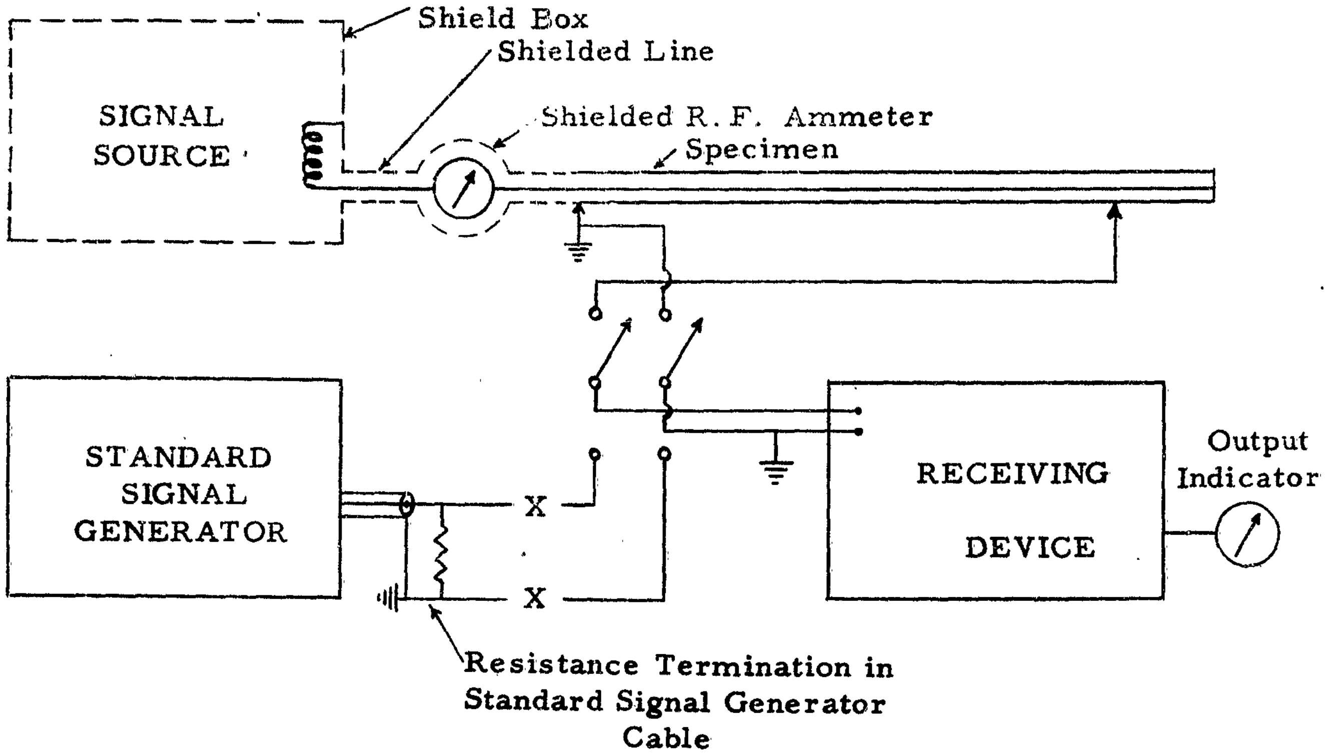


Fig. XI-2 Basic Circuit Used for Measuring Transfer Impedance

before, is divided by the current indicated on the ammeter yielding the desired transfer impedance.

2. Equipment Required and Suggested Types

(a) Oscillator-Amplifier Signal Source. TS-606/U and TS-608/U, Rollin Models 20 and 30, or equivalent. Care must be exercised in maintaining a good sinusoidal waveform, especially at the lower frequencies.

(b) RF Ammeter, Thermocouple Type. Weston, Model 640 or equivalent.

(c) Standard Signal Generator. Measurements Corporation Model 65-B or equivalent.

(d) A suitable receiving device. Ferris Meter or equivalent.

(e) A frequency standard, used to maintain signal source and generator within close tolerances. Millen Type 90501 or equivalent.

This method was used to investigate shielding effectiveness for frequencies up to 12 megacycles, and the range can probably be extended to several hundred megacycles.

Method II

1. Principle of Operation

The radio-frequency output of a signal generator is applied across the terminals of a radiating coil located within a completely shielded test cabinet, as indicated in

Figure XI-3. The coil is magnetically coupled to a probe connected directly to the input of a radio receiver. The output of the generator and the gain of the receiver are then adjusted to give a 10-milliwatt reading on an output power meter connected to the receiver output.

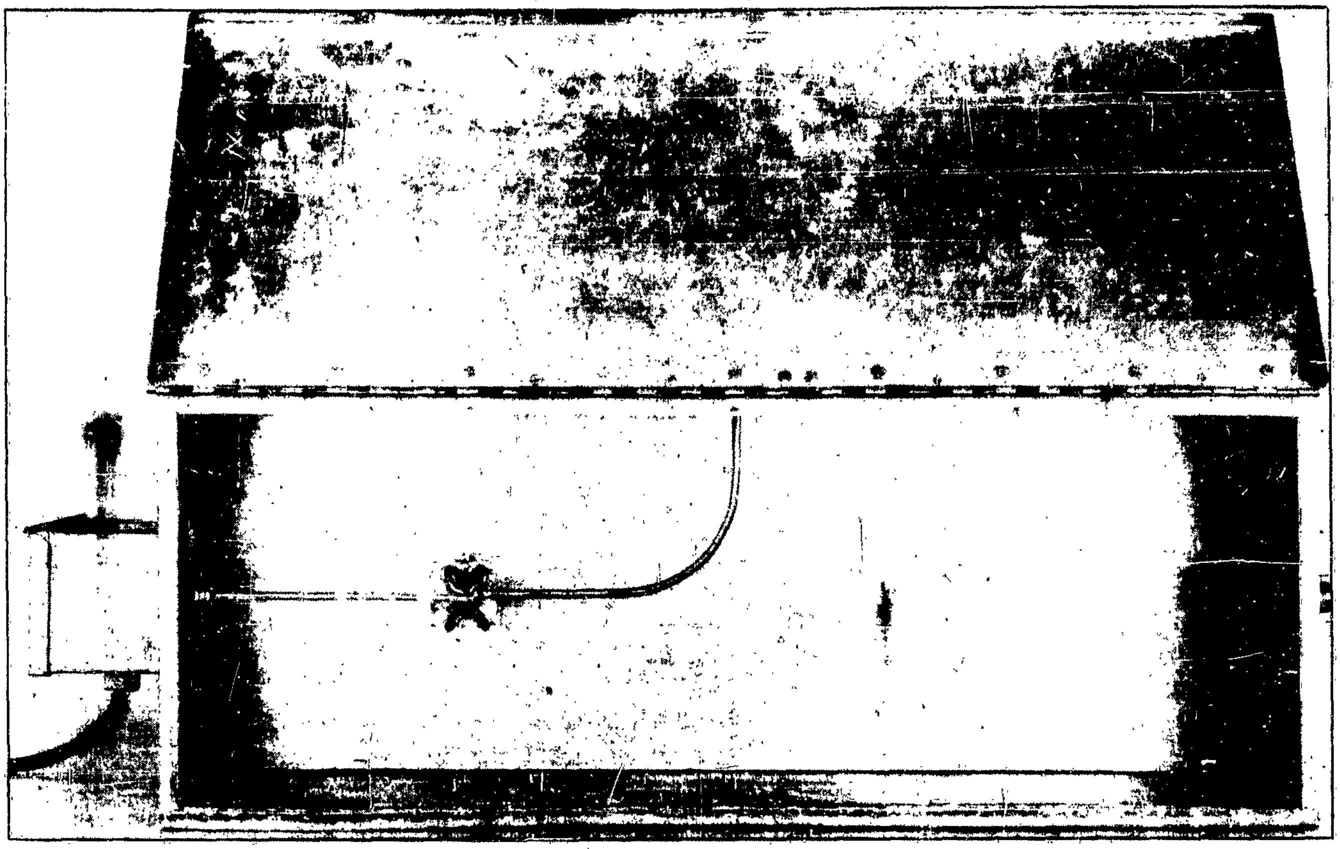


Fig. XI-3 Test Cabinet Illustrating Complete Shielding of Conduit Tester

A specimen of shield is then inserted between the radiating coil and probe. The signal generator only is adjusted to obtain the same output as before.

The ratio of generator voltage required with the shield for a given output to the voltage required without the shield is a measure of the shielding effectiveness.

2. Basic Test Set-Up

The equipment specified is suitable for measuring shielding effectiveness of electromagnetic field strength attenuation less than 100 db in the range 0.15 to 50 mc, and less than 45 db in the range of 50 to 156 mc. These limits may be extended by using signal generators of greater output or receivers of greater sensitivity.

In this frequency range, a typical radiating coil consists of seven turns of copper wire wound in the form of a solenoid, located within a ploystyrene sheath, and surrounded by a Faraday shield as shown in Figure XI-4. The Faraday shield insures repeatable measurements in the frequency range above 100mc.

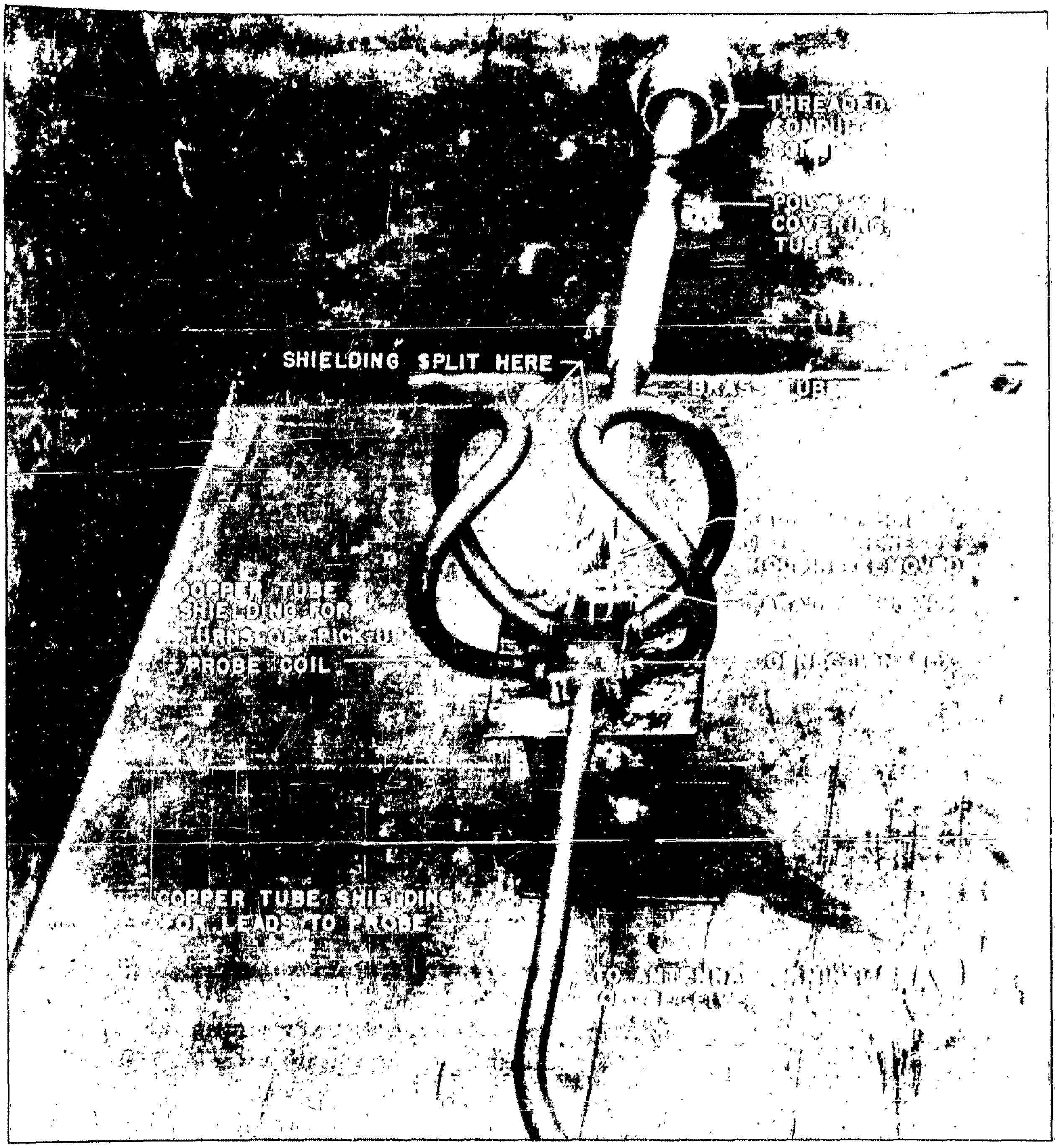


Fig. XI-4 Detail of Pick-Up Probe and Radiating Coil

The pick-up probe consists of two loops of copper tube shielding, within each of which is an insulated conductor. The loops are placed one on each side of the radiating coil in that position which permits linking the maximum number of flux lines in the field of the radiating coil. The loops are soldered into a small copper connection box at the base as shown in Figure XI-4. The copper tube shields the insulated conductor against electrostatic coupling, but in order to permit unimpaired magnetic coupling to the insulated conductor, the shielding is split at the top of each turn and the two cut ends are separated by a short air gap.

The pair of insulated conductors is continuously shielded to the rear wall of the test cabinet where one of them is connected to the inner surface of the tube that contains it, which is grounded to the wall of the cabinet. The other conductor is connected to the antenna post of the test receiver.

3. Equipment Required and Suggested Types

(a) Standard Signal Generator, General Radio Model 805-A or equivalent. (Range, 16 kc - 50 mc; maximum voltage output, 2 volts.)

(b) Standard Signal Generator. General Radio Model 804-C or equivalent. (Range, 50 - 156 mc; maximum voltage output, 0.02 volts. It is recommended that a signal generator of greater calibrated output be used if available.)

(c) Commercially available receivers (range to match signal generator). NC-200 and S-27 suggested.

(d) Output Power Meter. General Radio Type 583-A or equivalent.

4. Instructions for Test (See Figure XI-5)

(a) Place the test cabinet on a grounded metal test bench. Solder a strip of bonding braid (as short as possible) from each of the four corners of the test cabinet to the metal test bench.

(b) Connect the output of the signal generator to the coaxial cable connector mounted on the cabinet connection box.

(c) Connect a coaxial cable from the probe terminating connector on the rear of the test cabinet to the antenna post of the receiver. The cable should be bonded to the test bench at 18 inch intervals. (The antenna post should be shielded to prevent stray pick-up.) Connect the receiver ground post (or chassis) to the metal test bench with a short strip of bonding braid.

(d) Connect the output power meter to the receiver and adjust its impedance to correspond with the receiver output impedance.

(e) Connect the instruments to a 110 volt AC regulated power source. Allow a 1/2 hour warm-up period.

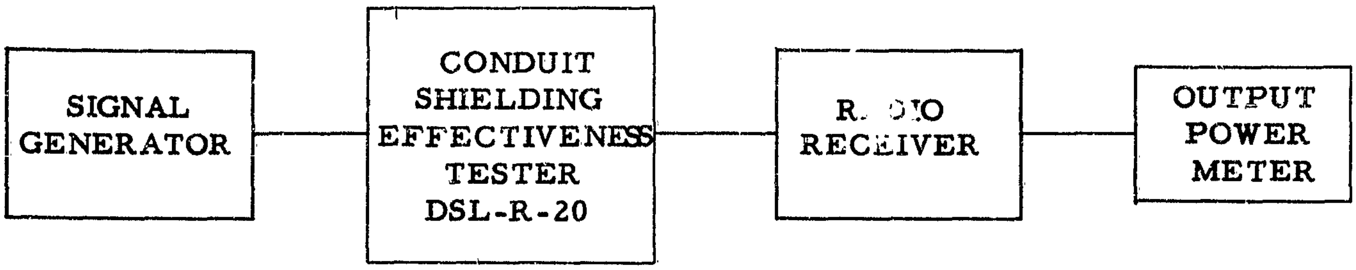


Fig. XI-5 Block Diagram for Shielding Effectiveness Test

5. Test Procedure

(a) With no conduit installed, but with the cabinet lid tightly closed, set the signal generator to the desired test frequency. Increase the output to enable detection of the signal.

(b) Tune the radio receiver to the same frequency, observing the output power meter for maximum indication. (Be sure that the receiver is not tuned to the "image" frequency.)

(c) Reduce the signal generator output to zero and adjust the receiver gain to obtain not more than two mw of residual noise.

(d) Increase the signal generator output until a 10 mw reading is obtained on the output power meter. Record the signal generator output as E_2.

(e) Install the conduit sample.

(f) Tightly close the cabinet lid.

(g) Increase the signal generator output until a 10 mw reading is again obtained on the output power meter. Record the signal generator output as E_1.

(h) Determine the ratio of the two signal generator output signals. Convert the ratio into decibels using the expression $20 \log_{10} (E_1/E_2)$. This figure represents the shielding effectiveness of the conduit.

APPENDIX XII

MEASURING THE RADIO-FREQUENCY IMPEDANCE OF BONDS

Insertion-loss measurements, which indirectly supply information from which the impedance may be computed, are made in preference to the direct measurement of radio frequency impedances because of the many difficulties encountered in the latter. The insertion-loss ratio is defined as the ratio of voltages existing across a load impedance before and after connecting the two-terminal test impedance in parallel with it. The insertion-loss ratio of any specimen under test is given by the expression:

$$\text{Insertion Loss Ratio} = \left| 1 + \frac{Z_s \, Z_R}{Z(Z_s + Z_R)} \right| \tag{1}$$

The quantity Z represents the impedance of the specimen under test, Z_s and Z_R represent the source and load impedances respectively, as shown in Figure XII-1.

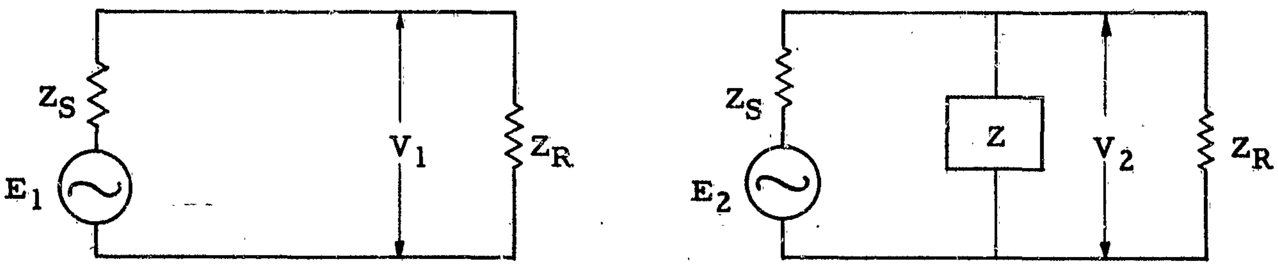


Fig. XII-1 Equivalent Circuits Used in the Derivation of the Insertion-Loss Ratio

The derivation of this expression can be obtained as follows. Let V_2 and V_1 represent the voltage drops across the load impedance with and without the specimen under test connected, and E_2 and E_1 represent the generator voltages in each case. Then:

$$V_1 = \frac{E_1 \, Z_R}{Z_s + Z_R} \tag{2}$$

$$V_2 = \frac{E_2 \, Z \, Z_R}{Z\,(Z_s + Z_R) + Z_s \, Z_R} \tag{3}$$

$$\frac{V_1 \, E_2}{V_2 \, E_1} = 1 + \frac{Z_s \, Z_R}{Z\,(Z_s + Z_R)} \tag{4}$$

When $E_2 = E_1$, i.e., generator voltage constant, the insertion loss is as defined above. It is seen that the same expression is obtained when the load voltage is kept constant, i.e., $V_1 = V_2$, and the generator voltage is adjusted accordingly. This second method is used in practical measurements.

In test set-ups used for measuring insertion loss, the source and load impedances are usually made resistive and equal to one another by means of isolation network pads. This simplifies the expression for the insertion-loss ratio which is now given as

$$\text{Insertion Loss Ratio} = \left|1 + \frac{R}{2Z}\right| \quad (5)$$

where R represents the equal resistive values. Furthermore, the resistive values are often arbitrarily fixed at 50 ohms which changes the expression to

$$\text{Insertion Loss Ratio} = \left|1 + \frac{25}{Z}\right| \quad (6)$$

It must be recognized that the insertion-loss ratio as measured between 50 ohm resistors will not represent the actual insertion loss of the impedance under test when used between the wide range of impedances encountered in practice. However, as long as some standard for source and load impedances is accepted and consistently used, test results are significant and comparison of data is valid since there is a direct correlation between insertion loss and impedance.

It is often convenient to use the insertion loss, measured in decibels, rather than the insertion loss ratio.

$$\text{Insertion Loss} = 20 \log_{10} \left|1 + \frac{R}{2Z}\right| \quad (7)$$

A circuit for measuring insertion loss is shown schematically in Figure XII-2.

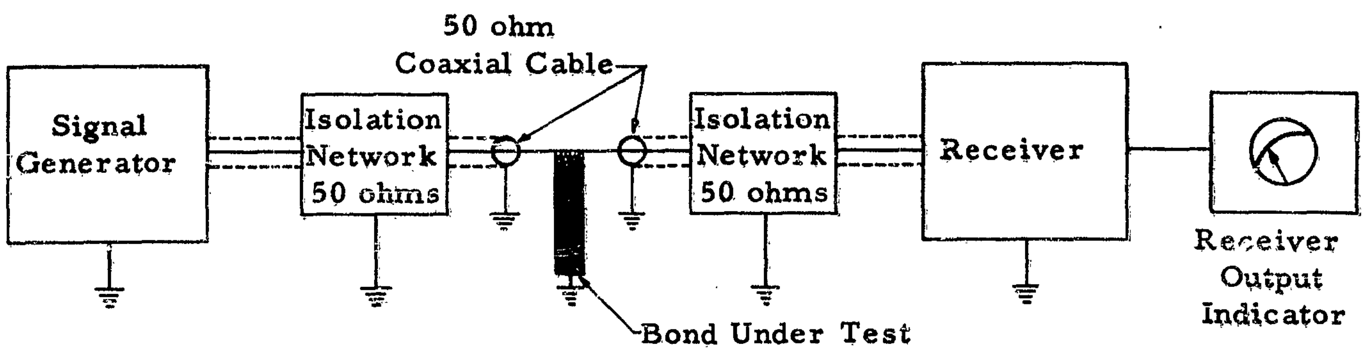


Fig. XII-2 Schematic Diagram for Measuring Insertion Loss

To make a measurement at any frequency the bonding jumper under test is removed from the circuit, a signal is introduced and the receiver output is recorded. The bonding jumper is then inserted into the circuit, and the signal generator output is raised until the same receiver output is obtained. The ratio of the second generator output reading to the first is the insertion loss ratio.

Several precautions must be taken in order to obtain accurate results. The length of the open line between the extremities of the coaxial cables must be minimized to prevent the introduction of an appreciable value of inductance in series with the source and load impedances. This is conveniently accomplished by connecting the extremity of each coaxial cable to a connector. Each of the two connectors is equipped with a stud no longer than 1/4 inch. These studs are connected directly when a measurement in the absence of a jumper is made. When a measurement is made with a bonding jumper present, they are firmly connected to the lug at the extremity of the jumper as shown in Figure XII-3.

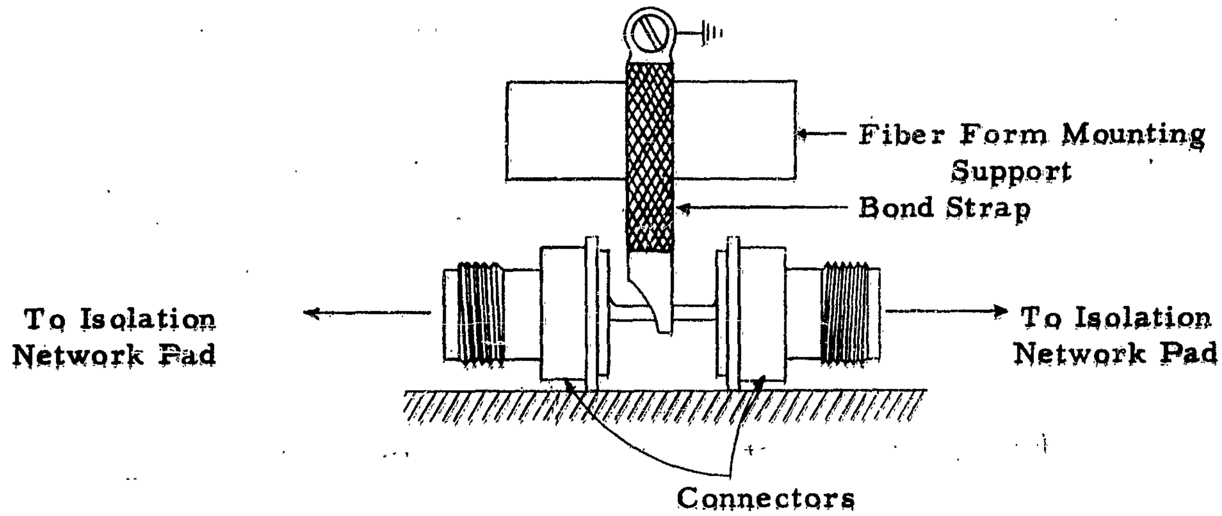


Fig. XII-3 Method of Connecting Bonding Jumper to Studs

The lug at the jumper's other extremity is securely bolted to the ground plate. A fixed orientation with respect to ground for all test samples must be ensured. This may be accomplished by securing the bonding jumper to a cylindrical, fiber mounting form as shown in Figure XII-4.

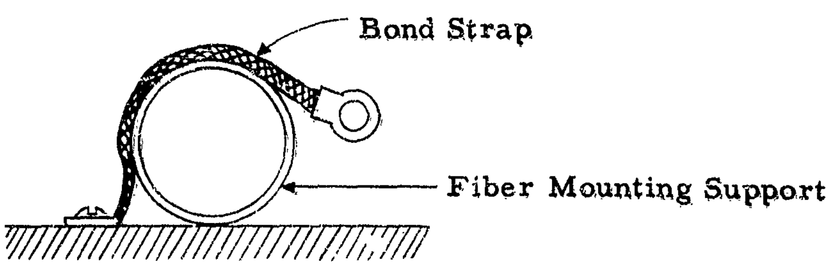


Fig. XII-4 Method of Connecting Bonding Jumper to Ground

Difficulties due to the impedance of the ground path at high frequencies can be avoided by mounting the signal generator, the isolation network pad on the source side of the circuit, and the specimen under test on a fiber plate. This isolates the source side from the receiver side and prevents the ground-path impedance from affecting the results. The return current path is, of course, through the outer conductor of the coaxial cable. In order to prevent a high capacity to ground, which results in a low impedance by-pass, the fiber plate must not be less than 1/4 inch thick.

APPENDIX XIII

DESCRIPTIVE DATA SHEETS OF INTERFERENCE TEST SETS

RADIO TEST SET AN/PRM-1

FUNCTIONAL DESCRIPTION:

A portable radio interference and field intensity measuring equipment used for radio interference surveys to determine the source of radiated or conducted interference from any source within its frequency range. It may also be used as a sensitive two-terminal voltmeter making both symmetrical and asymmetrical measurements with respect to ground such as measuring conducted interference on power and transmission lines. This set is used for field and depot operation. Frequency meter indicator and indicating meter are located on the front panel of the receiver.

Measurements can be made with the receiver in terms of the peak value of the interference (the PEAK function), in terms of a weighted value (the QUASI-PEAK function), or in terms of the average value (the FIELD INTENSITY function).

Signal monitoring provisions are available from oscilloscope, panel - mounted

(Continued)

	AIR FORCE	NAVY	ARMY
TYPE CLASS.	Approved	Approved	Approved
STOCK NOS.			
PROCUREM'T INFO.:			
PROCUREM'T COG.: Navy		DESIGN COG.: Navy, BuShips	
F. I. I. N.:		RDB IDENT. NO.: 5.5	
29 March 1952	- Electronics Test Equipment -		AN/PRM-1

FUNCTIONAL DESCRIPTION: (Continued)
headphones, and external meter receptacles.

RELATIONSHIP TO OTHER EQUIPMENT:
Similar to the Stoddart Model NM-20A.

ELECTROMECHANICAL DESCRIPTION:
Circuit Information: A seven band superheterodyne receiver is used with provision for attenuation and measurement of detector output.

The RF signal as picked up by the antenna or probe is delivered to the RF input. The RF signal is amplified in the RF stage and mixed with the local oscillator frequency in the mixer stage to produce an intermediate frequency. The IF signal is amplified in four IF stages and demodulated in the detector stage. The demodulated signal is acted upon by the meter detector weighting circuits and applied to the VTVM stage, thus actuating the meter. The audio components are subsequently amplified and delivered to the headphone jacks.

Power Supply: 115 volts, ±10%, AC, or 230 volts, ±10%, AC, single phase, 50 to 1600 cps, 25 watts. 9 volts supplied by 2 Batteries JAN-B-31 (4.5 volts) and 1.5 volts supplied by 2 Batteries JAN-58 (1.5 volts). The batteries required when AC is not available are 90 volts supplied by 2 Batteries JAN-BA-36 (45 volts) and 1.5 volts supplied by 2 Batteries JAN-BA-35 (1.5 volts).

Frequency Range: 0.15 to 25 megacycles per second in the following seven bands: 0.15 to 0.32, 0.32 to 0.75, 0.75 to 1.75, 1.75 to 3.8, 3.8 to 8, 8 to 15, 15 to 25 megacycles per second.

Intermediate Frequency Range: 455 kilocycles per second for bands 1, 3, and 4.
1600 kilocycles per second for bands 2, 5, 6, and 7.

Meter Scale: 0 to 100 microvolts, 0 to 40 db.

Voltage Range: 1 microvolt to 1 volt.

Field Intensity Range: 10 microvolts per meter to 100,000 microvolts per meter. (For Antenna AT-211/PRM-1).
1000 microvolts per meter to 10 volts per meter (For Antenna AT-212/PRM-1).
2 microvolts per meter to 2 volts per meter (For Antenna AT-213/PRM-1).

Selectivity: Overall bandwidth 3 to 5 kilocycles at 6 db down.
Overall bandwidth 20 to 30 kilocycles at 60 db down.

Audio Output: 100 milliwatts or better.

Audio Output Impedance: 600 ohms.

Dynamic Range: 16 db.

MANUFACTURERS' OR CONTRACTORS' DATA:
Stoddart Aircraft Radio Company, 6644 Santa Monica Boulevard, Hollywood 38, California, Contract No. NObsr-39262 dated 6/27/47; Contract No. NObsr-43370 dated 6/9/49.

TUBE COMPLEMENT:
IM-37/PRM-1: 1 JAN-1R5, 4 JAN-1T4, 1 JAN-1U5, 3 JAN-3A5, 4 JAN-3V4.
PP-472/PRM-1: 1 JAN-2A20, 1 JAN-0A3/VR75.

(Continued)

REFERENCE DATA AND LITERATURE:
Navships 91255 (Instruction Book).

SHIPPING DATA:

No. of Boxes	Contents & Identification	Volume (Cu. Ft.)	Over-all Dimensions (inches)			Weight Packed (Lbs.)
			H	W	D	
1	Radio Test Set, AN/PRM-1	8.0	21-1/2	28	23	141 approx.

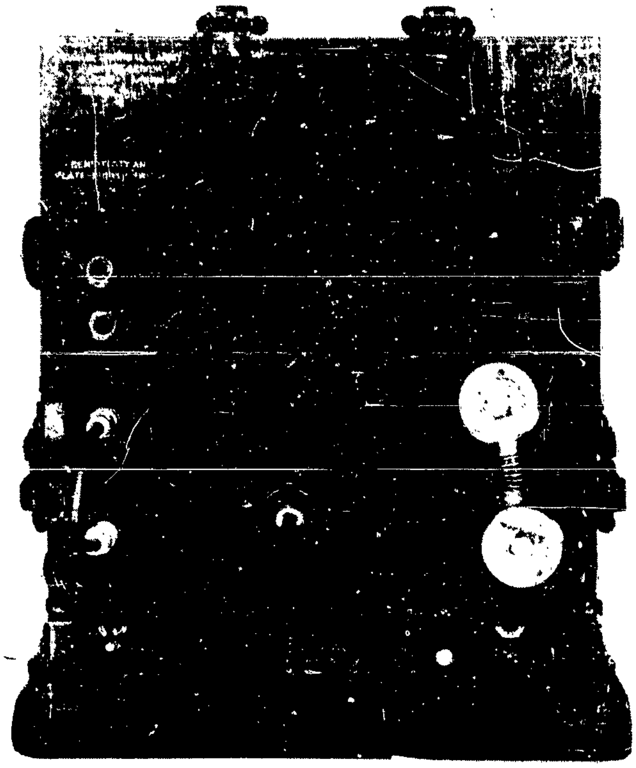
EQUIPMENT SUPPLIED:

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches)			Weight (Lbs.)
				H	W	D	
1	Radio Interference Field Intensity Meter IM-37/PRM-1			14	7-9/16	16-7/16	37
1	Transit Case CY-749/PRM-1			18-3/4	19-1/4	11-3/8	17
1	Antenna AT-211/PRM-1						
1	Antenna AT-212/PRM-1			8-1/4 dia.			
1	Antenna AT-213/PRM-1			41 long			
1	RF Probe MX-980/PRM-1						
1	Impedance Matching Network CU-195/PRM-1						
1	Impedance Matching Network CU-196/PRM-1						
1	Impedance Matching Network CU-197/PRM-1						
1	Adapter UG-104/U						

EQUIPMENT SUPPLIED: (Continued)

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches)			Weight (Lbs.)
				H	W	D	
1	Adapter UG-105/U						
1	Adapter UG-537/U						
1	Special Purpose Cable Assembly CG-572/U			240 long			
1	Chart Set PT-117/PRM-1						
1	Headphone CW-49509						
2	Instruction Book Nav-ships 91255						
1	Shoulder Strap						
1	Accessory Case CY-750/PRM-1			18-3/8	15-5/8	10-3/4	16
1	Power Supply PP-472A/PRM-1			7-17/32	10-1/2	8-3/8	18
1	Ammeter ME-33/U						
1	Power Cable Assembly CADV-62480			72 long			
1	Special Purpose Cable Assembly CADV-62481			120 long			
1	Special Purpose Cable Assembly CG-571/U			240 long			
1	Cord CG-444/U			240 long			
1	Special Purpose Cable Assembly CG-573/U			36 long			

TEST SET AN/URM-3



FUNCTIONAL DESCRIPTION:

A general purpose, field, and maintenance equipment used for detecting and measuring the intensity of radiated and conducted radio interference. Probes are provided for conducting exploratory interference tests and matching and coupling networks are used to permit it to be used as a two-terminal RF microvoltmeter.

RELATIONSHIP TO OTHER EQUIPMENT:

ELECTROMECHANICAL DESCRIPTION:

Circuit Information: Consists basically of a superheterodyne receiver and a calibrated impulse noise generator generating pulses exhibiting a stable and uniform spectrum throughout the range of the receiver, the peak value of which is adjustable to known values. The receiver is a six-band receiver with two tuned RF stages, a mixer, local oscillator, two IF stages, second detector, first audio and output stages. The output of the impulse generator is injected into the receiver antenna input circuit in such a manner that interference measurement is independent of antenna impedance.

(Continued)

	AIR FORCE	NAVY	ARMY
TYPE CLASS:	Accepted		Accepted
STOCK NOS.			
PROCUREM'T INFO.: Signal Corps, Exhibit A, 6/24/46.			
PROCUREM'T COG.: Army		DESIGN COG.: Army, CSL	
F.I.I.N.:		RDB IDENT. NO.: 5.5	
3 April 1952	- Electronics Test Equipment -		AN/URM-3

ELECTROMECHANICAL DESCRIPTION: (Continued)

Power Supply: 115 volts, ±10%, AC, single phase, 60 cycles per second; or 24 volts DC; or 12 volts DC.

Frequency Range: 0.15 to 0.4 and 1.6 to 40 megacycles per second in six bands. The ranges are 0.15 to 0.4, 1.6 to 3.0, 3.0 to 5.8, 5.8 to 11.0, 11.0 to 21.0, and 21.0 to 40.0 megacycles per second.

Voltage Range: 10 microvolts per megacycle to 31,600 microvolts per megacycle.

Radio Noise Generator:

Pulse Duration: Approximately 0.01 microsecond.

Pulse Repetition Rate: 10 to 1000 pulses per second.

Pulse Amplitude: 0 to 90 db above one microvolt per megacycle of bandwidth.

Spectrum: Flat to 40 megacycles per second within ±3 db.

MANUFACTURERS' OR CONTRACTORS' DATA:

Designed by the Signal Corps Engineering Laboratories, Fort Monmouth, New Jersey.

TUBE COMPLEMENT:

R-178/URM-3: 1 JAN-0B2/VR105, 1 JAN-0A3/VR75, 1 JAN-0C3/VR105, 1 JAN-6AC7W, 1 JAN-6H6, 2 JAN-6SA7Y, 1 JAN-6SG7Y, 2 JAN-6SK7W, 1 JAN-6SQ7, 1 JAN-6V6GT.

TS-496/URM-3: 2 JAN-0A2, 1 JAN-0B3, 1 JAN-6J6, 1 JAN-5696.

REFERENCE DATA AND LITERATURE:

Preliminary Instruction Manual, 9/1/50, Signal Corps Engineering Laboratories, Fort Monmouth, New Jersey.

SHIPPING DATA:

No. of Boxes	Contents & Identification	Volume (Cu. Ft.)	Over-all Dimensions (inches)			Weight Packed (Lbs.)
			H	W	D	

EQUIPMENT SUPPLIED:

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches)			Weight (Lbs.)
				H	W	D	
1	Test Set AN/URM-3			24-1/2	12	14-1/2	110
1	Mast Base AB-15/GR						
1	Antenna Bracket MT-195/URM-3						

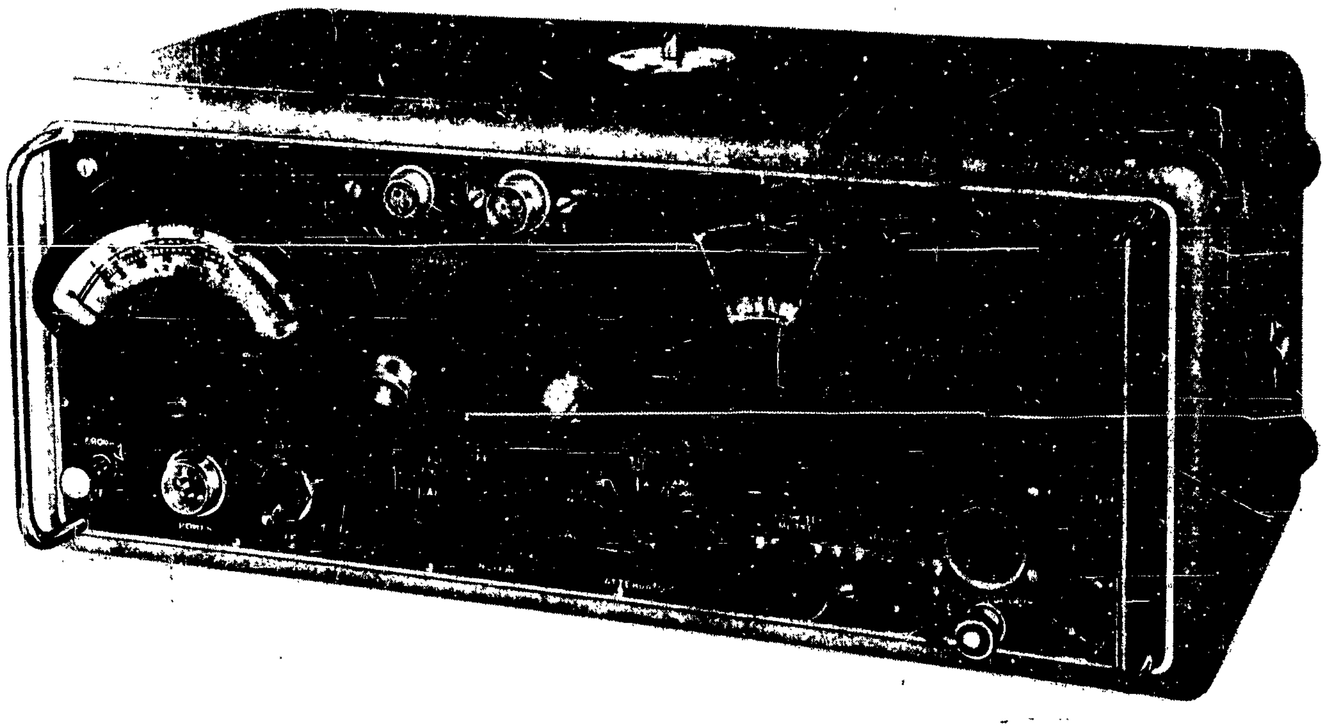
EQUIPMENT SUPPLIED: (Continued)

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches) H	W	D	Weight (Lbs.)
1	Cable Assembly RG-55/U			180 long			
1	Cable Assembly RG-55/U			60 long			
1	Adapter UG-273/U						
1	Headset H-16/U						
1	Cord (Headset) CD-307						
1	Probe, Electric Field MX-839/URM-3						
1	Probe, Magnetic Field MX-838/URM-3						
1	Mast Section MS-116						
1	Mast Section MS-117						
1	Mast Section MS-118						
1	Cord (DC Power) CX-1063/URM-3						
1	Cord (AC Power) CX-1062/URM-3						
1	Coupler CU-149						
1	Coupler CU-150						
1	Coupler CU-153						

EQUIPMENT SUPPLIED: (Continued)

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches)			Weight (Lbs.)
				H	W	D	
1	Coupler CU-151						
1	Coupler CU-152						
1	Adapter Connector UG-641/U						
1	Service Patch Cord						

RADIO TEST SET AN/URM-6



FUNCTIONAL DESCRIPTION:

A field and depot equipment used to measure the field intensity of a given radio transmission, to measure the intensity of radiated or conducted radio interference, or as a sensitive RF microvoltmeter within its frequency range. Two tripods are provided for mounting the large loop antenna and the superheterodyne receiver. Mounting provisions have been made on the receiver for either of two rod antennas and a small loop antenna. Signal monitoring provisions are available from panel-mounted headphones, oscilloscope, recorder, and remote meter jacks. Measurements are made using the panel-mounted meter or the remote meter and graphic recordings are made using the milliammeter-recorder. Input devices include rod antennas, loop antennas, a line probe, and impedance matching networks. A calibration chart is furnished with the equipment.

Measurements can be made with the receiver in terms of the peak value of the signal or interference (the PEAK function), in terms of the nuisance value (the QUASI-PEAK function), or in terms of the average value (the FIELD INTENSITY function).

RELATIONSHIP TO OTHER EQUIPMENT:

Similar to Stoddart Model NM-10A.

(Continued)

	AIR FORCE	NAVY	ARMY
TYPE CLASS.	Approved	Approved	Approved
STOCK NOS.			
PROCUREM'T INFO.:			
PROCUREM'T COG.: Navy		DESIGN COG.: Navy, BuShips	
F.I.I.N.:		RDB IDENT. NO.: 5.5	
31 March 1952	- Electronics Test Equipment -		AN/URM-6

ELECTROMECHANICAL DESCRIPTION:

Circuit Information: The power input goes through an isolation transformer with filters to keep extraneous power-line noises out of the instrument. Meter indications must be modified by the pick-up factor of the antenna used. The receiver is a highly sensitive low frequency radio receiver which contains internal means for calibrating or standardizing its RF gain, thus permitting direct readings in indicated microvolts or microvolts per meter.

The signal channel of the receiver resembles a conventional superheterodyne receiver in its RF, IF, and AF portions, but differs in its provision for attenuation and measurement of detector output. The RF signal is amplified in the RF stage and mixed with the local oscillator frequency in the mixer stage to produce an intermediate frequency. The IF signal is amplified in three IF stages and demodulated in the detector stage. The demodulated signal is acted upon by the meter detector weighting circuits and applied to the VTVM stage. The audio components are subsequently amplified and delivered to the headphone jacks.

Attenuator step ratio settings are built in the inputs of the RF, mixer, and IF stages.

Power Supply: 115 volts, ±10%, or 230 volts, ±10%, AC, single phase, 50 to 1600 cycles per second except for Milliammeter-Recorder which utilizes 60 cycles per second, 100 watts at 115 volts, 60 cycles. 3.0 volts DC, supplied by two Battery, BA-30, for the Observer-Compass, Mark 1, Model O. A suitable battery pack can be used in place of the separate power supply, however, the Milliammeter-Recorder cannot be used.

Frequency Range: 14 to 250 kilocycles per second.

Intermediate Frequency: 12.5 kilocycles per second.

Audio Output: 100 milliwatts.

Audio Impedance: 600 ohms (headset).

Attenuator Setting: 0, 20, 40, 60, and 80 db.

Receiver Meter Scale: 0-100 microvolts, 0-40 db.

Voltage Range: 1 microvolt to 1 volt.

Field Intensity Range: 1 microvolt per meter to more than 1 volt per meter, depending on the antenna used.

Effective Bandwidth: 100 cycles to 600 cycles at 6 db down, 2000 cycles at 60 db down.

Image Rejection: -50 db or better from signal level.

Intermediate Frequency Rejection: 60 db or better.

Signal-to-Noise Ratio: Unity or better.

Dynamic Range: 20 db at full scale.

Accuracy: Field Intensity Measurements, ±10% above 10 microvolts per meter.

MANUFACTURERS' OR CONTRACTORS' DATA:

Stoddart Aircraft Radio Company, 6644 Santa Monica Boulevard, Hollywood 38, California, Contract No. NObsr-39263, 6/27/47.

TUBE COMPLEMENT:

IM-36/URM-6: 1 JAN-6AL5, 1 JAN-6AT6, 6 JAN-6AU6, 1 JAN-6BE6, 3 JAN-6C4, 1 JAN-6E5, 1 JAN-6J6, 1 JAN-6X4, 1 JAN-NE2.

PP-449/URM-6: 1 JAN-5V3GT, 1 JAN-NE32, 2 JAN-0C3/VR-105.

REFERENCE DATA AND LITERATURE:

Navships 91196 (Instruction Book).

SHIPPING DATA:

No. of Boxes	Contents & Identification	Volume (Cu. Ft.)	Over-all Dimensions (inches)			Weight Packed (Lbs.)
			H	W	D	
1	Radio Test Set AN/URM-6	26.2	25-1/2	37-1/2	47-1/2	297

EQUIPMENT SUPPLIED:

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches)			Weight (Lbs.)
				H	W	D	
1	Radio Interference Field Intensity Meter IM-36/URM-6			8	19-13/16	10-3/8	26
1	Transit Case CY-706/URM-6	Plywood		14-1/2	10-7/8	34-1/2	30
1	Power Supply PP-449/URM-6			7-31/32	9-29/32	19-3/4	24.75
1	Antenna AT-203/URM-6			51 long			
1	Antenna AT-204/URM-6			84 long			
1	Antenna AT-205/URM-6			5 dia.			
1	Adapter UG-537/U						
1	RF Cable Assembly CG-577/URM-6						
1	Power Cable Assembly CADV-62480			72 long			
1	Special Purpose Cable Assembly CADV-62481			120 long			
1	Chart PT-107/URM-6						
1	Clipboard						

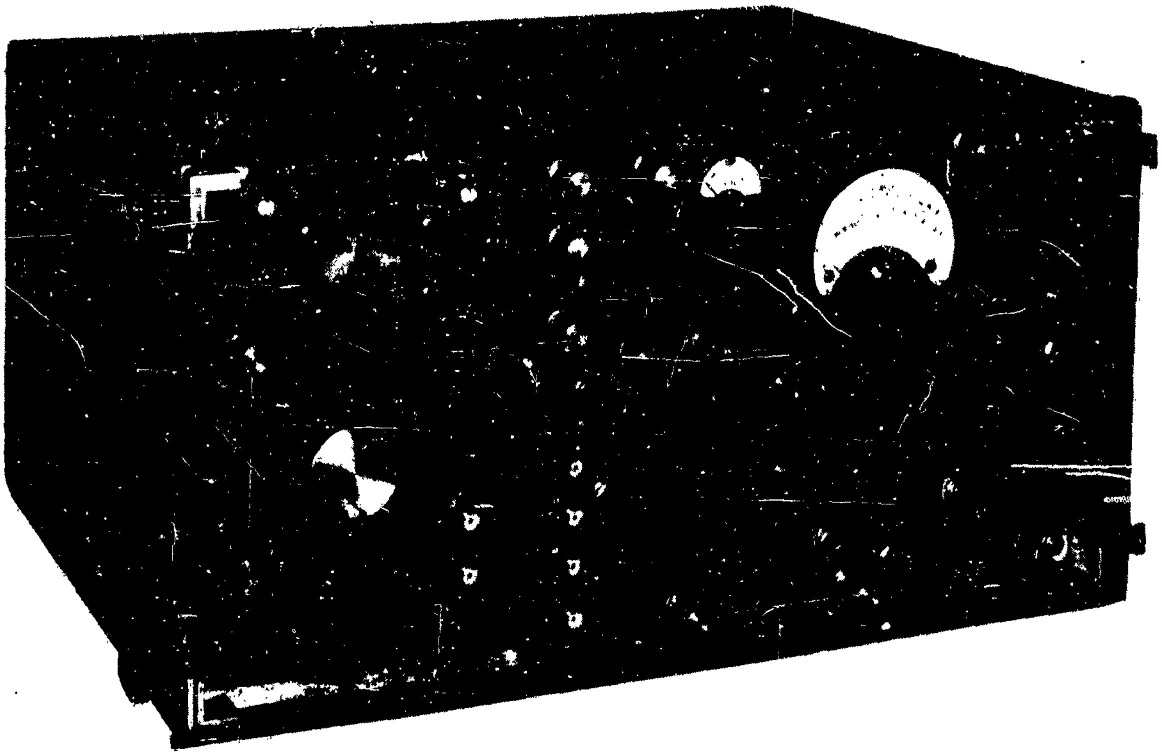
EQUIPMENT SUPPLIED: (Continued)

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches) H	W	D	Weight (Lbs.)
2	Instruction Book Navships 91196						
1	Headphone CW-49509						
1	Accessory Case CY-707/URM-6	Plywood		21-1/8	6-3/4	27-3/4	24
1	Observer-Compass with case Mark 1, Model O			4-3/4	10-7/8	4-7/8	4.75
1	Ammeter ME-31/U						
1	Antenna AT-207/URM-6						
1	RF Probe MX-951/URM-6						
1	Impedance Matching Network CU-184/URM-6						
1	Impedance Matching Network CU-185/URM-6						
1	Impedance Matching Network CU-186/URM-6						
1	Cord CG-444/U			240 long			
1	Special Purpose Cable Assembly CG-571/U			240 long			
1	Special Purpose Cable Assembly CG-572/U			240 long			
1	Special Purpose Cable Assembly CG-571/U			72 long			

EQUIPMENT SUPPLIED: (Continued)

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches) H	W	D	Weight (Lbs.)
1	Special Purpose Cable Assembly CADV-62482			24 long			
1	Special Purpose Cable Assembly CG-573/U			36 long			
1	Case (Recorder) CY-708/URM-6	Ply-wood		18	14	11	15
1	Milliammeter-Recorder with inking kit RD-59/U			13-1/4	8-9/16	8-3/4	
1	Power Cable Assembly CADV-62480			72 long			
1	Shoulder Strap						
1	Tripod Case CY-709/URM-6	Can-vas		41-1/2	9-1/4 dia.		5
2	Tripod MT-674/U						
1	Loop Case CY-710/URM-6	Ply-wood		33-3/4	5	33-7/8	36
1	Antenna AT-206/URM-6			30 dia.			
1	Shoulder Strap						
						Total:	222.0

RADIO INTERFERENCE MEASURING SET, AN/URM-7



FUNCTIONAL DESCRIPTION:

A general purpose, field radio interference and field intensity meter designed primarily for the measurement of broadband interference although it incorporates facilities for CW interference and field intensity measurement. Test set incorporates an impulse generator used as a noise reference standard, whose output is calibrated in terms of microvolts per unit bandwidth. The visual output indicator is a peak reading vacuum tube voltmeter with a logarithmic scale calibrated in microvolts and a linear decibel scale calibrated in terms of decibels above one microvolt per megacycle. Probes are provided for conducting exploratory interference tests, and coupling networks are used to permit the Test Set to be used as a two-terminal noise-microvoltmeter.

RELATIONSHIP TO OTHER EQUIPMENT:

AN/URM-7 is similar to Empire Device, Inc., commercial Model NF-105.

ELECTROMECHANICAL DESCRIPTION:

Circuit Information: The impulse generator output is injected into the input circuit of the tuner in such a manner as to permit measurement of open circuited antenna

(Continued)

	AIR FORCE	NAVY	ARMY
TYPE CLASS.			
STOCK NOS.			
PROCUREM'T INFO.:			
PROCUREM'T COG.: Army		DESIGN COG.: Army, CSL	
F.I.I.N.:		RDB IDENT. NO.: 5.6	
27 February 1952	- Electronics Test Equipment -		AN/URM-7

ELECTROMECHANICAL DESCRIPTION: (Continued)

terminal voltage on a per megacycle basis. The tuner utilizes a superheterodyne circuit. The frequency range is covered by means of two plug-in type RF heads. The logarithmic scale characteristic of the output indicator is achieved through tapered pole-pieces in the indicating meter movement, which eliminates the necessity of using automatic gain control. The dipole antenna, used for field intensity measurement applications, can be resonated at each test frequency. The broadband antenna is used in suppression test applications where the antenna must be placed close to the source of interference.

Power Supply: 115 volts, ±10%, AC, single phase, 50 to 400 cycles, 100 volt-amperes; or 24 volts, DC; or 12 volts, DC.

Frequency Range: 20 to 200 mc, 200 to 400 mc.

Intermediate Frequency Range: 10.7 mc, 30 mc.

Voltage Range: 12 µv/mc to 1,200,000 µv/mc (for 20 to 200 mc).
6 µv/mc to 5,000,000 µv/mc (for 200 to 400 mc).

Indicating Meter Scale: 0.5 to 10 µv.
-6 to +20 db (10 db scale expansion is provided for scale overlap).

Calibration Standards:

(a) Spot frequency sine wave generator.
(b) Broadband impulse noise generator (output externally available).
Pulse duration: 5×10^{-4} microseconds.
Pulse Repetition Rate: 2.5 to 2500 pulses per second.
Pulse Amplitude: 47 to 97 db above one microvolt per megacycle bandwidth.
Spectrum: Flat to 1000 mc within ±1/2 db.

Accuracy: ±10%, voltage.

MANUFACTURERS' OR CONTRACTORS' DATA:

Empire Devices, Inc., 38-25 Bell Boulevard, Bayside, New York, Contract No. W36-039-sc-38120.

TUBE COMPLEMENT:

8 JAN-6BJ6, 4 JAN-12AT7, 3 JAN-12AU7, 5 JAN-6X4, 1 JAN-0A2, 5 JAN-6AK5, 1 JAN-6AL5, 1 JAN-6AB4, 1 JAN-6J6, 1 JAN-6F4, 1 JAN-5876, 1 JAN-1N21B (Crystal), 1 JAN-1N34 (Crystal).

REFERENCE DATA AND LITERATURE:

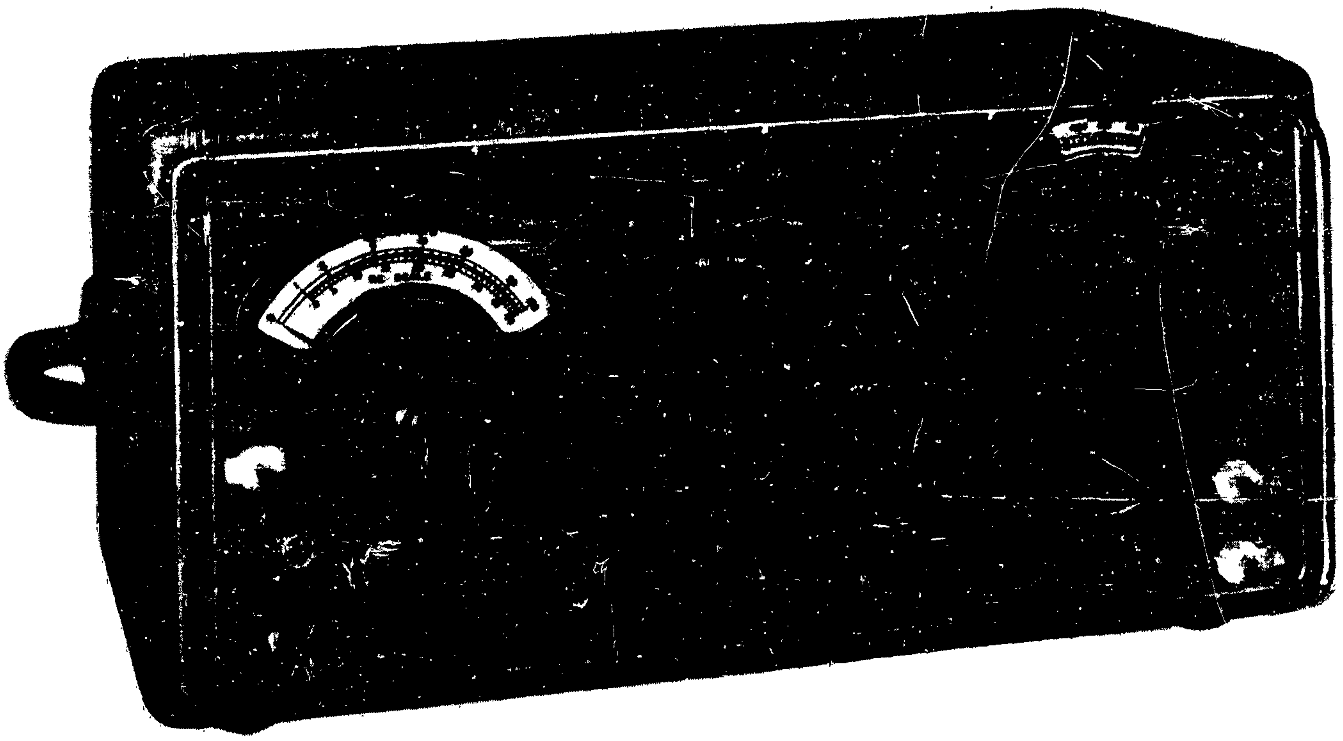
SHIPPING DATA:

No. of Boxes	Contents & Identification	Volume (Cu. Ft.)	Over-all Dimensions (inches)			Weight Packed (Lbs.)
			H	W	D	

EQUIPMENT SUPPLIED:

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches) H	W	D	Weight (Lbs.)
1	Radio Interference Measuring Set, AN/URM-7			9-1/2	18-3/4	14-5/8	65
1	Carrying Case (Spare)						
1	Accessory Carrying Case						
1	Dipole Antenna Set						
1	Tripod						
1	Line Probe (50 ohms)						
1	Magnetic-Field Probe						
1	Electric-Field Probe						
1	Balanced Injection Block						
1	Unbalanced Injection Block						
2	RF Cable			360 long			
1	Power Line Cord						
1	Earphone						
1	Alignment Harness						
1	Operation Manual						

RADIO TEST SET AN/URM-17



FUNCTIONAL DESCRIPTION:

A field and depot equipment used for intensity measurements of all type of radio-frequency energy in the radio frequency spectrum. It contains internal means for calibrating or standardizing its gain. Measurements can be made with the receiver in terms of the peak value of the interference (the PEAK function), in terms of a weighted value (the QUASI-PEAK function), or in terms of the average value (the FIELD INTENSITY function). The indicating meter scale on the panel is directly calibrated for a two decade, approximately logarithmic range of 1-100 microvolts, and an approximately linear 0-40 db range. Charts provided with the equipment show actual effective bandwidth versus frequency.

RELATIONSHIP TO OTHER EQUIPMENT:

Similar to Stoddart Model NM-50A.

Equipment required but not supplied: One Headphone, Navy type CW-49509 or equivalent; one Observer Compass, AN designation Mark 1, Model O.

ELECTROMECHANICAL DESCRIPTION:

Circuit Information: The signal channel closely resembles a superheterodyne receiver in its RF, IF, and AF portions, but differs from most superheterodyne

(Continued)

	AIR FORCE	NAVY	ARMY
TYPE CLASS.			
STOCK NOS.			
PROCUREM'T INFO.:			
PROCUREM'T COG.: Navy		DESIGN COG.: Navy, BuShips	
F.I.I.N.:		RDB IDENT. NO.: 5.5	
2 April 1952	- Electronics Test Equipment -		AN/URM-17

ELECTROMECHANICAL DESCRIPTION: (Continued)

receivers in its provision for attenuation and measurement of detector output.

The RF signal as picked up by the antenna or probe is delivered to the RF Input receptacle and passes through the RF stage. It is then mixed with the local oscillator frequency in the mixer stage, to produce the intermediate frequency. The IF signal is amplified in six IF stages and demodulated in the detector stage. The demodulated signal is acted upon by the meter detector weighting circuits and applied to the VTVM stage, thus actuating the meter. The audio components are subsequently amplified and delivered to the headphone jacks.

Power Supply: 115 volts, ±10%, AC, or 230 volts, ±10%, AC, single phase, 50 to 1600 cycles per second, 110 watts at 115 volts or a suitable battery pack used in place of the separate power supply.

Frequency Range: 375 to 1000 megacycles per second.

Intermediate Frequency: 60 megacycles per second.

Input Impedance: 50 ohms.

Attenuator Steps: 60 db to 140 db.

Effective Bandwidth: Approximately 1.8 megacycles at signal of 1000 megacycles to 1.0 megacycles at signal of 370 megacycles.

Image Rejection: 40 db or better.

Spurious Response Rejection: 40 db or better.

Intermediate Frequency Rejection: Better than 60 db.

Voltage Range: 100 microvolts to 10 volts.

Field Intensity Range: 100 microvolts per meter to 100 volts per meter.

Signal-to-Noise Ratio: Unity or better based on an equipment sensitivity of 10 microvolts as a two-terminal voltmeter.

Audio Output: 100 milliwatts or better.

Audio Output Impedance: 600 ohms.

Dynamic Range: 20 db.

MANUFACTURERS' OR CONTRACTORS' DATA:

Stoddart Aircraft Radio Company, 6644 Santa Monica Boulevard, Hollywood 38, California, Contract No. NObsr-42430, 6/30/48.

TUBE COMPLEMENT:

IM-52/URM-17: 2 JAN-6AL5, 1 JAN-6AR5, 7 RMA-6BAG, 2 JAN-6C4, 2 JAN-6F4, 1 JAN-9005, 1 JAN-12AU7, 1 JAN-1N21B (Crystal Rectifier).

PP-530/URM-17: 1 JAN-6AS7G, 1 RMA-6BH6, 1 JAN-NE-32.

REFERENCE DATA AND LITERATURE:

Navships 91388 (Instruction Book).

SHIPPING DATA:

No. of Boxes	Contents & Identification	Volume (Cu. Ft.)	Over-all Dimensions (inches)			Weight Packed (Lbs.)
			H	W	D	
1	Radio Test Set, AN/URM-17	8.3	18-1/4	43-1/2	19	170

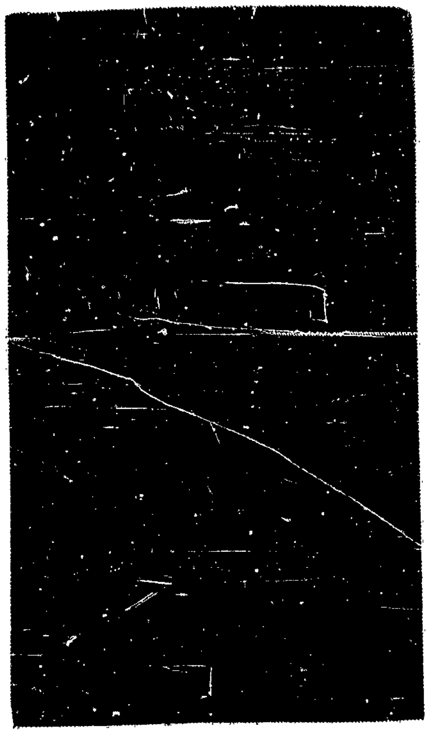
EQUIPMENT SUPPLIED:

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock (USAF) Numbers (Navy) (Army)	Over-all Dimensions (inches) H	W	D	Weight (Lbs.)
1	Radio Interference Field Intensity Meter IM-52/URM-17			9-3/16	19-13/16	10-3/8	30
1	Transit Case CY-866/URM-17	Plywood		16-1/4	23-3/4	10-7/8	25
1	Chart Set PT-210/URM-17						
2	Instruction Book Navships 91388						
1	Accessory Case CY-865/URM-17	Plywood		14-5/8	15-1/2	10-1/8	17
1	Power Supply PP-530/URM-17			7-31/32	9-29/32	19-3/4	16
1	Antenna AT-255/URM-17	Brass	F16-A-45194-7901	5-1/2 (collapsed) 16 (extended)			
1	RF Probe DT-56/URM-17		N16-R87008-4021	2-31/32 long 9/16 OD			
1	Impedance Matching Network CU-227/URM-17		N16-N30826-1073	2-31/32 long 9/16 OD			
1	Power Cable Assembly CADV-62480			72 long			
1	Special Purpose Cable Assembly CADV-62481		N17-C-48703-5521	120 long			
1	RF Cable Assembly CG-678/U		N16-C-11957-3021	240 long			
1	Glow Lamp JAN-NE-32 (Spare)		2Z5889-8				
1	Ballast Lamp (Spare)		N16-R-85001-1511				

EQUIPMENT SUPPLIED: (Continued)

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches)			Weight (Lbs.)
				H	W	D	
1	Cable Connector (20 ohms)			1-11/16 long	53/64 OD		
1	Tripod Bag CW-218/URM-17	Can-vas		39-3/4 high	6-1/4 dia.		3.25
1	Tripod Navy Type-10545	Wood	N16-T-802001-107	37-1/2 long (collapsed) 60 long (extended)			
1	Antenna Mast Section AB-189/URM-17			36-1/8 long	1 dia.		
1	Azimuth Dial Assembly	Alum-inum	N16-I-21801-1045	3 long	2-1/4 dia.		
1	Azimuth Dial Pointer		N16-K-701645-401	3/4 long	2 dia.		
						Total:	109.25

RADIO NOISE AND FIELD STRENGTH METER
FERRIS MODEL 32-B



FUNCTIONAL DESCRIPTION:

A portable field, depot interference and field strength meter used for measurements of carrier voltages and fields, noise voltages and fields, signal-to-noise ratio voltages on lines and conductors, antenna field patterns, filter characteristics, localization of noise sources and interference reduction means, and all kinds of radio interference fields.

The tuning dial is graduated in megacycles, while the output meter scale reads directly in microvolts the value of the voltage applied to the input.

RELATIONSHIP TO OTHER EQUIPMENT:

Similar to Model 32-A, Radio Noise and Field Strength Meter, except Model 32-B has increased sensitivity and a five microvolt output meter scale.

ELECTROMECHANICAL DESCRIPTION:

Circuit Information: A single RF stage feeds into a combination mixer and oscillator tube. Two IF stages precede the detector tube but the AF stage is omitted. The circuits are designed for compactness and simplicity of operation.

(Continued)

	AIR FORCE	NAVY	ARMY
TYPE CLASS.	Accepted		
STOCK NOS.	Commercial		
PROCUREM'T INFO.:			
PROCUREM'T COG.:		DESIGN COG.: Commercial	
F.I.I.N.:		RDB IDENT. NO.: 5.6	
29 March 1952	- Electronics Test Equipment -		Model 32-B

ELECTROMECHANICAL DESCRIPTION: (Continued)

Power Supply: Dry cell large battery pack (32BP2). Small battery pack, AC battery pack, or receifier packs may be obtained on special request.

Frequency Range: 150 to 350 kc and 500 to 20,000 kc in five bands.

Voltage Range: 0-5, 1000, 100,000 microvolts.

Field Intensity Range: 1 to 200,000 microvolts (With standard Antenna).

Standard Time Constant: Charge, 10 milliseconds; discharge, 600 milliseconds.

Sensitivity: 0.5 microvolts.

Calibrating Source: Internal shot noise oscillator (Calibration curves are supplied).

MANUFACTURERS' OR CONTRACTORS' DATA:

Designed and manufactured by Ferris Instrument Company, Boonton, New Jersey,

TUBE COMPLEMENT:

4 RMA-1D5GP, 1 RMA-1C7G, 1 JAN-1H6G, 1 JAN-1H4G, 1 RMA-1G1 (Ballast Tube).

REFERENCE DATA AND LITERATURE:

Ferris Radio Noise and Field Strength Meter Model 32-B (Operating Instructions).

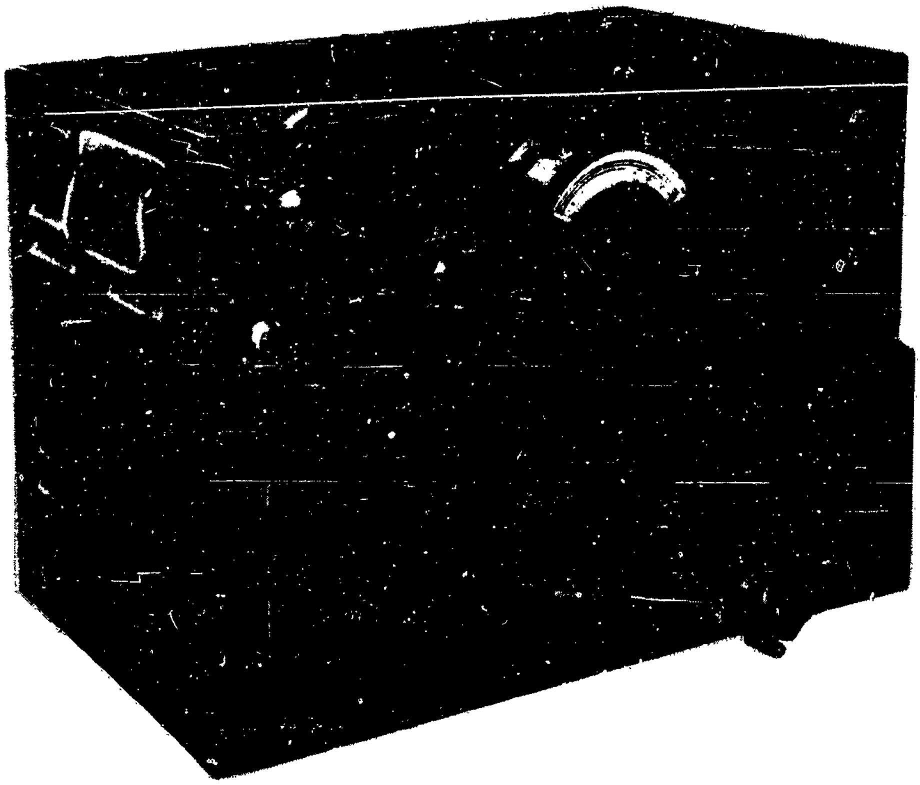
SHIPPING DATA:

No. of Boxes	Contents & Identification	Volume (Cu. Ft.)	Over-all Dimensions (inches)			Weight Packed (Lbs.)
			H	W	D	
1	Model 32-B					38

EQUIPMENT SUPPLIED:

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches)			Weight (Lbs.)
				H	W	D	
1	Model 32-B Including Cover		Commercial	13-1/4	7	14-1/2	32
1	Telescoping Rod Antenna			41 long			

NOISE-FIELD INTENSITY METER TS-587A/U



FUNCTIONAL DESCRIPTION:

A field and depot equipment used to locate and measure RF interference and to make field strength measurements. It can be used as a two terminal sensitive voltmeter. A tripod is furnished for mounting the antennas. Signal monitoring provisions are available from the phone jacks. Measurements can be made with the receiver in terms of the peak value of the interference (the PEAK function), in terms of a weighted value (the QUASI-PEAK function), or in terms of the average value (the FIELD INTENSITY function). The measurements are made using the panel-mounted meter or the remote meter jacks. The meter scale is directly calibrated in microvolts, decibels, and arbitrary units of "shot noise". Arbitrary units of "shot noise" are used only in the calibration procedures of the equipment. By use of correction curve furnished with the equipment, the meter readings can be converted to microvolts per meter.

RELATIONSHIP TO OTHER EQUIPMENT:

TS-587A/U differs from TS-587/U only in minor mechanical and electrical details that do not affect its operational characteristics or servicing.

Similar to the Stoddart Model NMA-5A.

(Continued)

	AIR FORCE	NAVY	ARMY
TYPE CLASS.	Accepted	Approved	Accepted
STOCK NOS.			
PROCUREM'T INFO.:			
PROCUREM'T COG.: Navy		DESIGN COG.: Navy, BuShips	
F.I.I.N.:		RDB IDENT. NO.: 5.5	
2 April 1952	- Electronics Test Equipment -		TS-587A/U

RELATIONSHIP TO OTHER EQUIPMENT: (Continued)

Modification Kit MX-910/U, consisting of a DC Amplifier and Rotary Converter, is used with TS-587A/U when suitable power is not otherwise available.

ELECTROMECHANICAL DESCRIPTION:

Circuit Information: It is a high-sensitivity HF and VHF superheterodyne radio receiver which contains internal means for calibrating its VTVM section. Four decade steps of attenuation are inserted between the antenna input and the RF head. The output signal of the attenuator is fed to the RF head, where it is amplified and mixed with the local oscillator frequency, producing the intermediate frequency. The IF signal is amplified in four IF stages and demodulated in the detector stage. The demodulated signal is acted upon by the meter detector weighting circuits and applied to the VTVM stage, thus actuating the meter. The audio components are subsequently amplified and delivered to the headphone jacks.

Power Supply: 115 volts, ±10%, AC, single phase, 60 cycles per second, 0.96 amperes at 92% power factor.

Frequency Range: 15 to 400 megacycles per second in four bands.
LF Head: 15 to 31, 29 to 64, 60 to 125 megacycles per second.
HF Head: 100 to 400 megacycles per second.

Intermediate Frequency: LF Head: 12 megacycles per second.
HF Head: 30 megacycles per second.

Voltage Scale: 0 to 100 microvolts.
-6 to 40 decibels.
0 to 8 arbitrary units of shot noise.

Input Voltage Range: 2 to 100,000 microvolts. Balance resistance attenuator with steps of: X1, X10, X10^2, and X10^3.

Receiver Output: 200 milliwatts maximum.

Input Impedance: 95 ohms, balance to ground.

Output Impedance: 300 or 4000 ohms, for headphone.

Sensitivity: Two-terminal Voltmeter: LF Head: 2 microvolts.
HF Head: 5 microvolts.
Field Intensity Meter: LF Head: 20 microvolts per meter.
HF Head: 5 microvolts per meter.

Bandwidth: LF Head: 150 kilocycles at 6 db down.
HF Head: 210 kilocycles at 6 db down.

MANUFACTURERS' OR CONTRACTORS' DATA:

Stoddart Aircraft Radio Company, 6644 Santa Monica Boulevard, Hollywood 38, California, Contract No. NObsr-30088 dated 6/15/46, Contract No. NObsr-30140 dated 6/21/40, Contract No. NObsr-30200 dated 6/26/46.

TUBE COMPLEMENT:

RF-36/U: 2 JAN-6AK5, 2 JAN-6C4.
RF-37/U: 2 JAN-6J4, 1 JAN-6J6, 1 JAN-9002.
AM-194/U: 1 JAN-6AL5, 1 JAN-6AQ6, 4 JAN-6BA6.
AM-195/U: 1 JAN-6AL5, 1 JAN-6AQ6, 4 JAN-6SG7.
PP-267/U: 1 JAN-0D3/VR-150, 1 JAN-5Y3GT/G, 1 JAN-6H6, 1 JAN-6J5GT/G, 1 JAN-6L6GA, 1 JAN-6SJ7, 2 JAN-6V6 or 6V6GT/G.

(Continued)

REFERENCE DATA AND LITERATURE:
Navships 900, 990 (Instruction Book).

SHIPPING DATA:

No. of Boxes	Contents & Identification	Volume (Cu. Ft.)	Over-all Dimensions (inches)			Weight Packed (Lbs.)
			H	W	D	
1	Noise-Field Intensity Meter TS-587A/U	35.3	36	47	36	330

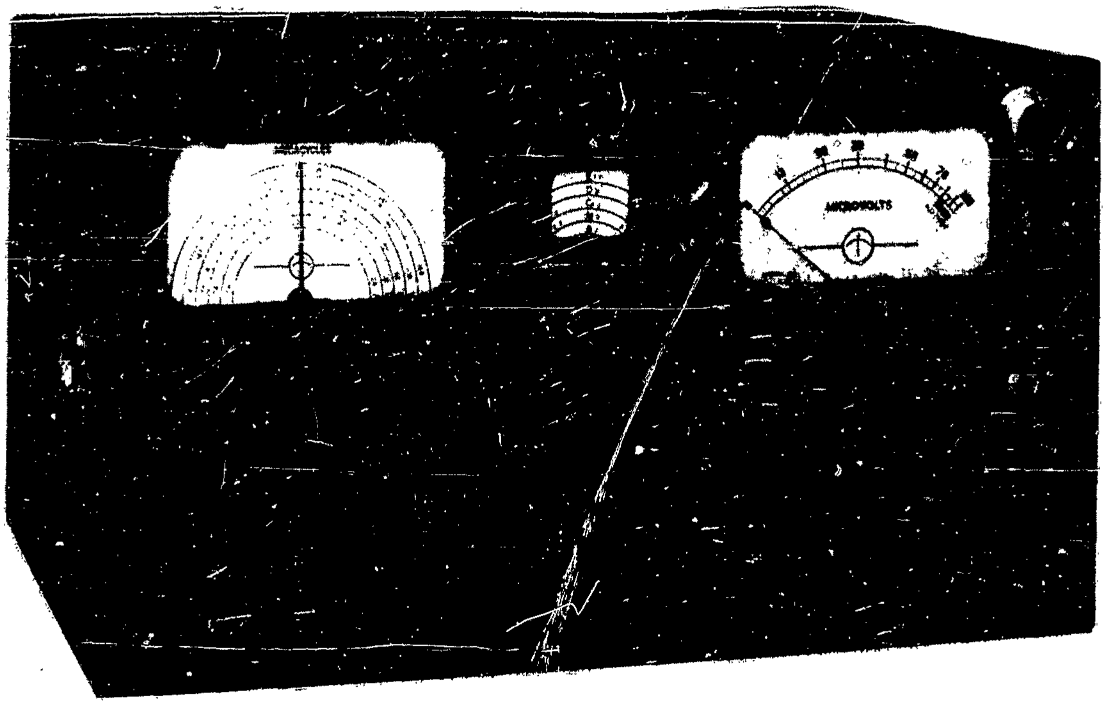
EQUIPMENT SUPPLIED:

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches)			Weight (Lbs.)
				H	W	D	
1	Noise-Field Intensity Meter TS-587A/U			15-3/16	21-1/2	29-1/2	64
1	Canvas Cover						
1	Equipment Case CY-606/U Containing:			15	40	9	23
1	Radio Frequency Head RF-36/U						
1	Intermediate Frequency Amplifier AM-194/U						
1	Dipole Antenna AS-385/U						
2	Dipole Mast Section for AS-385/U						
1	Loop Probe MX-822/U						
1	Measuring Tape CADV-10671						
2	Instruction Book Navships 900, 990						
1	Accessory Case CY-607/U			15	40	9	25

EQUIPMENT SUPPLIED: (Continued)

Quant. Per	Name and Nomenclature	Case Mat'l	Stock (USAF) Numbers (Navy)	Over-all Dimensions (inches) H	W	D	Weight Packed (Lbs.)
1	Dipole Antenna CADV-66151-A						
2	Dipole Mast Section for CADV-66151A						
1	Loop Probe MX-821/U						
1	Line Probe CADV-491087						
1	Antenna Tripod CADV-10545						
1	Matching Impedance CADV-471228						
1	Input Cable CG-44/U			72 long			
1	Power Cable						

MODEL 58 UHF RADIO NOISE AND FIELD STRENGTH METER



FUNCTIONAL DESCRIPTION:

A portable meter for measuring field strength of carrier voltages, for measuring ignition or diathermy interference, for determining front-to-back ratios of antennas, the effectiveness of noise filters, etc. It may also be used as a sensitive voltmeter for measuring RF voltage, transmission loss of various four-terminal networks, and for determining the interference level on power lines, battery leads, etc.

Frequency of incoming signal and value of received signal in microvolts are directly indicated.

Suitable jacks are provided for connecting an oscillograph, or a recording milliammeter to the audio frequency output.

A "slide-back" feature can be incorporated in the instrument, when desired, to facilitate the measurement of peak noise.

RELATIONSHIP TO OTHER EQUIPMENT:

Not specifically related to other equipment, but is used to test a wide variety of electronic and electrical devices for radiated radio frequency noise level.

(Continued)

	AIR FORCE	NAVY	ARMY
TYPE CLASS.	Accepted		
STOCK NOS.			
PROCUREM'T INFO.:			
PROCUREM'T COG.:		DESIGN COG.: Commercial	
F.I.I.N.:		RDB IDENT. NO.: 5.6	

ELECTROMECHANICAL DESCRIPTION:

Circuit Information: It is a five band superheterodyne receiver with its gain standardized at a fixed value. Four decade steps of attenuation are inserted between the antenna input and the first RF stage. The output of the detector tube excites the grid of an amplifier tube which is in one arm of a bridge circuit. The indicating microvoltmeter serves as a balance indicator across the bridge arm with readings that are semi-logarithmic.

Power Supply: 115 volts, ±10%, AC, single phase, 50 to 60 cycles per second, approximately 70 watts; or 6 volts, DC, 12 amperes.

Frequency Range: 15 to 150 megacycles in five bands: 15 mc-24 mc, 24 mc-39 mc, 38 mc-62 mc, 60 mc-100 mc, and 98 mc-150 mc.

Frequency Accuracy: Dials directly calibrated in megacycles to an accuracy of ±2%.

Input Voltage Range: 1 to 100,000 microvolts in the antenna. 1 to 100 microvolts on semi-logarithmic output meter, balanced resistance attenuator with steps of: X1, X10, X10^2, and X10^3.

Gain Standardization: Internal "shot" noise diode provides calibration standard. Special dial eliminates need for charts.

Band Width: 150 kilocycles at 6 db down.

MANUFACTURERS' OR CONTRACTORS' DATA:

Designed and manufactured by Measurements Corporation, Boonton, N. J.

TUBE COMPLEMENT:

3 JAN-7W7, 1 JAN-6X5GT, 1 JAN-6J5GT, 1 JAN-6Y6G, 1 JAN-6H6, 1 JAN-6SJ7, 1 JAN-955, 2 JAN-6AK5, 1 JAN-M74, 1 JAN-991 (Neon).

REFERENCE DATA AND LITERATURE:

Measurements UHF Radio Noise and Field Strength Meter Model 58 (Operating Instructions).

SHIPPING DATA:

No. of Boxes	Contents & Identification	Volume (Cu. Ft.)	Over-all Dimensions (inches)			Weight Packed (Lbs.)
			H	W	D	
1	Model 58 (Domestic Packed), (Export Packed).	2.9 3.7	14 17	22 22	16 17	53 66

EQUIPMENT SUPPLIED:

Quant Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches)			Weight (Lbs.)
				H	W	D	
1	Model 58			9	16	11	35
1	Loop Antenna A110			9 dia.			0.3

EQUIPMENT SUPPLIED: (Continued)

Quant. Per Eq'pt	Name and Nomenclature	Case Mat'l	Stock Numbers (USAF) (Navy) (Army)	Over-all Dimensions (inches)			Weight (Lbs.)
				H	W	D	
1	Antenna Cable A112			180 long			1.3
1	Power Cable (AC) A107			120 long			0.7
1	Power Cable (DC) A108			72 long			1.3

APPENDIX XIV

DETAILS OF LAMINATED BRUSH DESIGN

The use of laminated brushes for improving resistance commutation is explained in Paragraph 3.2.1.1.2. The details for constructing laminated brushes are given in this appendix.

Either a mixture of copper and bakelite or a mixture of graphite and bakelite is satisfactory for preparing brush stocks of different resistivity. The materials selected are ground to powder consistency, passed through a 297 micron sieve to assure particles of uniform size, and mixed in a ball mill in various proportions to obtain various degrees of resistivity. Figures XIV-A and B are curves giving the required mixing data necessary to predetermine resistivity with respect to percentage of mixture. The mixture is then placed in molds in a hydraulic press and baked for one hour at a temperature of 190° C. under a pressure of 2500 pounds per square inch. Laminations of different thicknesses are cut from the brush stock after removal from the mold. The laminations selected to fabricate a brush are coated with six mil leaf glue (No. R612 Minnesota Mining and Manufacturing Company is suitable) and are baked for one hour at a temperature of 180° under a pressure of 500 pounds per square inch. To maintain a constant temperature during the baking process, the oven is covered with a glass cloth and a variac is used to control the heat lamp employed. The brush thus formed is sanded to its exact dimensions, painted on both sides with G. E. 7031 adhesive, and baked at a moderate temperature to dry the adhesive. Cables are now tamped into the brush top, and to insure electrical contact across one end of the laminations, the brush top is subjected to a copper spray.

The techniques described above have been found satisfactory for constructing laminated brushes. Previously it was impossible to construct brushes having satisfactory radio interference characteristics because of the intermingling of the conducting and insulating materials. This was caused by the formation of minute pockets in the insulation between laminations which become filled with the conducting dust of the laminations due to brush wear. The gluing process described above and the use of a superior glue which can be subjected to relatively high pressures with a minimum of extrusion overcame this difficulty. Furthermore, brushes constructed as described possess satisfactory mechanical characteristics.

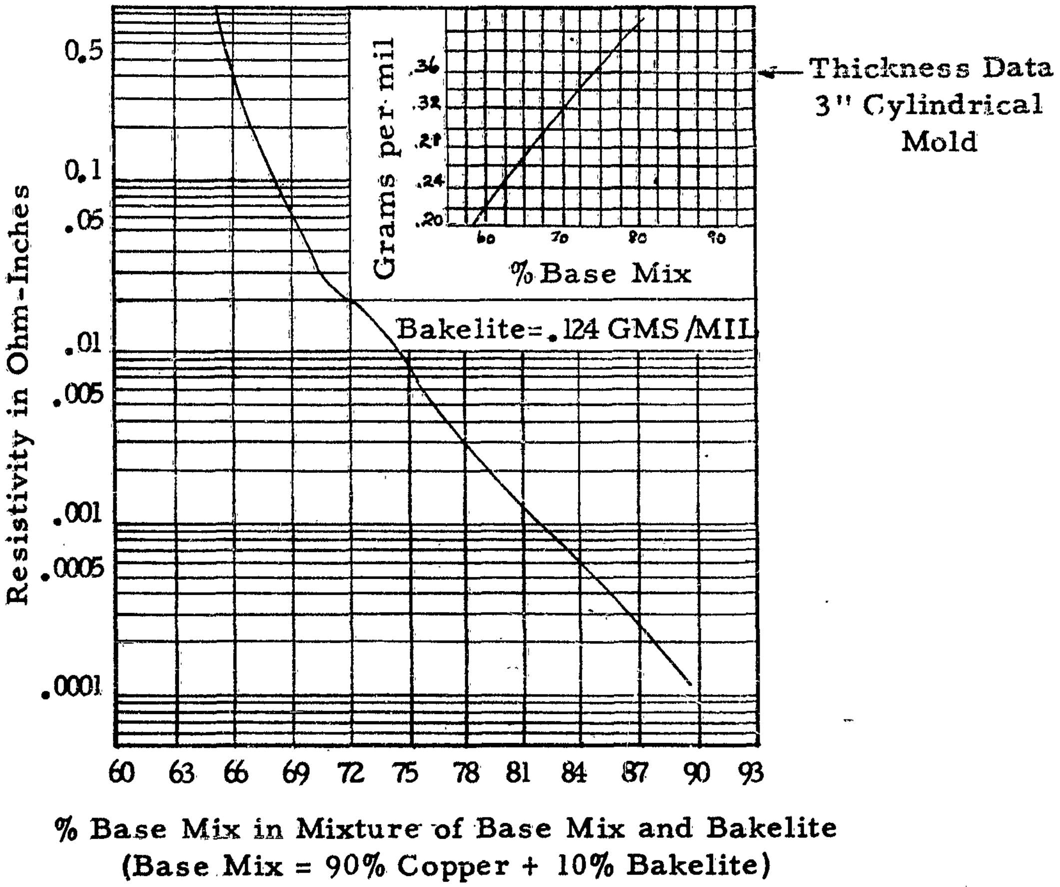


Fig. XIV-A Mixing Data For Copper Base Material

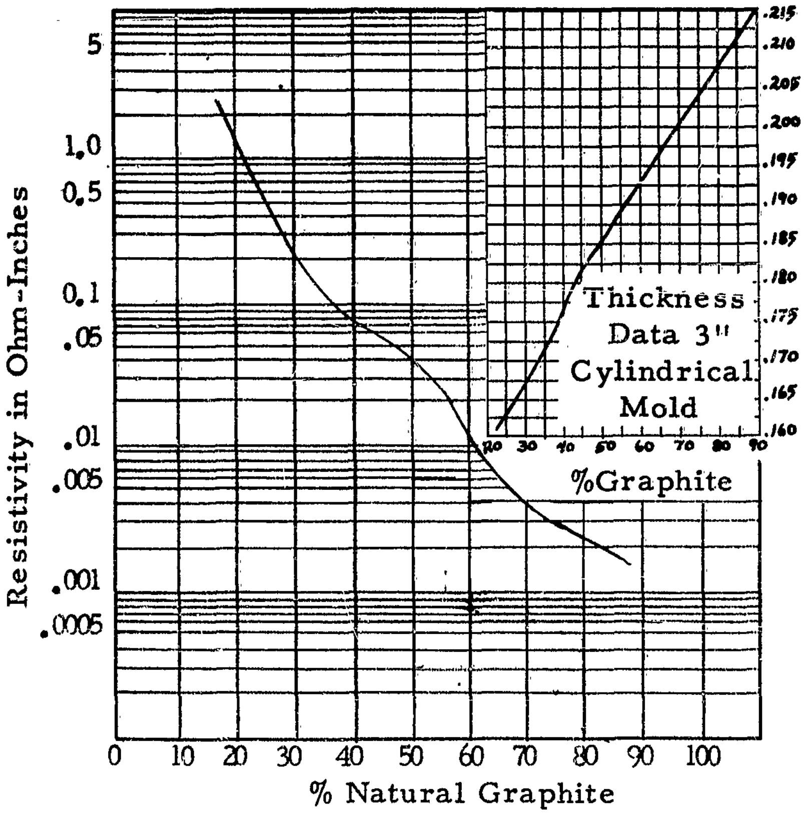


Fig. XIV-B Mixing Data For Graphite Bakelite

APPENDIX XV

CONSTRUCTION OF BONDING JUMPERS FOR SHOCKMOUNTED EQUIPMENT

An investigation into the best type of bonding jumper for shockmounted equipment has led to the development of a new design, the essential features of which are described in this appendix.

The requirements for a good bonding jumper are the following:

(a) Low direct current resistance

(b) Low radio frequency impedance at all frequencies up to 1000 mc

(c) Good mechanical properties as follows:

1. Ability to withstand ambient conditions and endurance requirements

2. Minimum height so that the overall mounting height of the shockmount is not increased

3. Absence of, or adequate guards against, all sharp edges for personnel protection

4. Minimum weight

5. Minimum volume

6. Minimum stiffness so that the vibration-isolation characteristics of the shockmount are not adversely affected.

Consideration of all these requirements has led to the development of the types illustrated in Figure XV. The most important feature is the employment of two metallic strips whose width is large compared to their thickness.

These jumpers consist of three major parts: the two bonding strips, the base plate, and the spacer washers. The strips are displaced 90° in position and connected externally between the top and bottom of the shockmount. Four strips could be used, but the addition of two more strips decreases the impedance by only ten percent and adds considerably to the damping. Therefore, it is not recommended.

A possible alloy to be used in the construction of the strips is Ni-Span "C" whose inherent properties are those of high modulus of elasticity, high tensile strength, and a high degree of hardness.

To limit the stress on the bonding strip its length should be made as great as possible while its thickness should be kept to a minimum. However, the electrical requirements of the bonding connector limit the length of the strips while ease in manufacture and fragility of relatively thin strips limit their thickness. For shockmounts

whose load capacities are one pound or less, a thickness of 0.002 of an inch is recommended; for mounts whose load capacities exceed one pound, a thickness of 0.003 of an inch should be used. A good compromise for the other dimensions is a length of from 1 to 3 inches and a width of from 3/4 inches to 1 inch.

The sharp edges of the thin leaves are safety-shielded with glass fibre sleeving, a material that will withstand the extreme ambient conditions expected, for the protection of personnel. The sleeving is coated with Dow-Corning No. 801 cement, and partially cured in an oven at approximately 300° F. Following this the strip is cut into sections of desired length, slipped over the bonding leaves, re-cemented, and cured completely.

The baseplate to which the bonding strips are welded is made of cadmium-plated SAE 1030 sheet steel. Where possible, the base is designed for top mounting because this design does not add appreciably to the overall mounting height of the shockmount.

Both spacer washers are made of cadmium-plated cold-rolled steel. The thickness of the upper spacer washer is sufficient to prevent contact between the bonding leaves and the plate which is to be shockmounted. Because of the possibility of excessive vibration the bottom spacer washer is made sufficiently large to function as a snubber.

Tortional stresses resulting from sideway motion of the top of the mount with respect to the base will be lessened by connecting the tops of the bonding leaves to the spacer washers by a loosely stacked joint. This joint lowers the calculated stresses (developed on the basis of a doubly clamped beam), and furthermore allows the rotation of the leaves plus slight lateral movements.

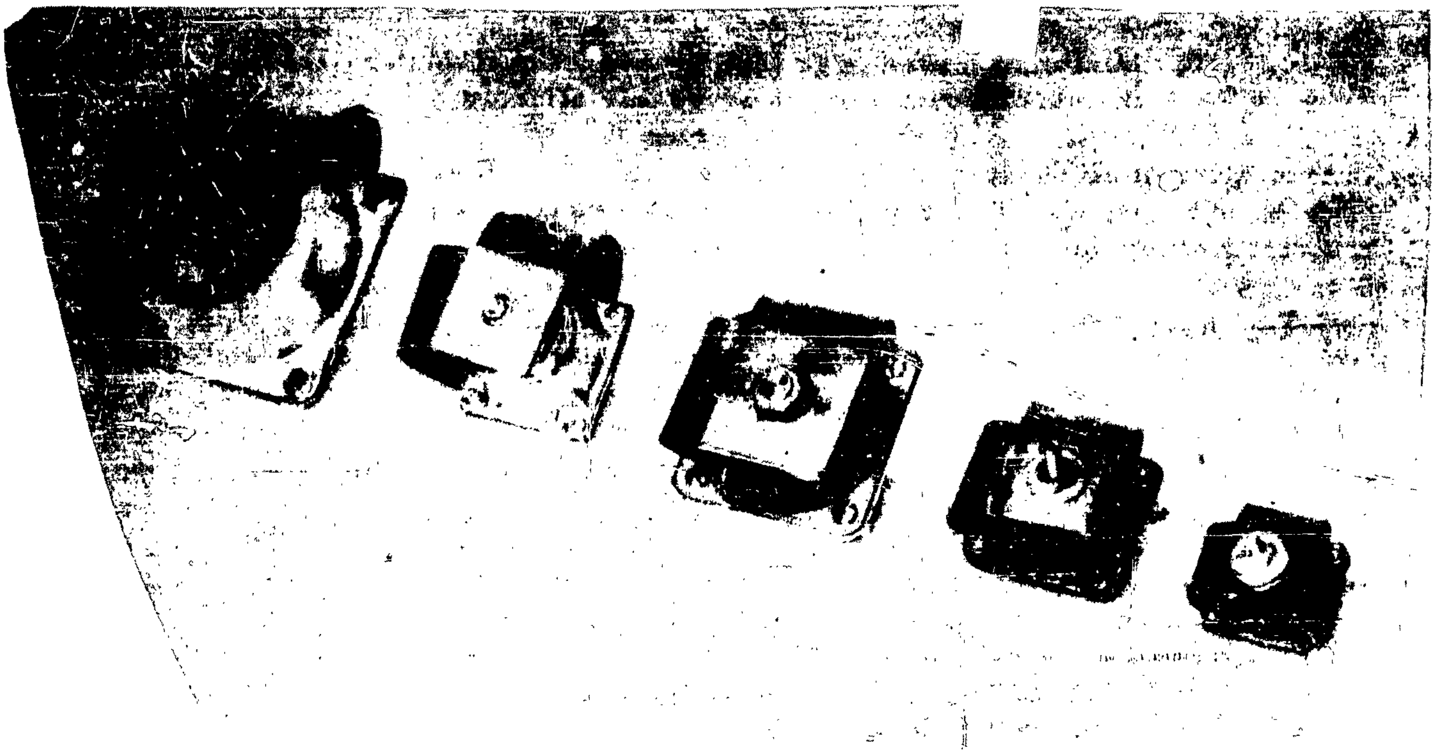


Fig. XV Capacitive Types Incorporated in Shockmount

The pin type joint at the top of the mount, designed with as thin sections as practical to reduce overall height, also reduces tortional stresses in the leaves. This joint is so designed that a positive electrical contact is assured, yet allows rotational movement without binding.

The direct current resistance of the types shown varied from 0.012 to 0.059 ohms depending upon the length of the jumper, and was not seriously affected by the mutual movements of the strips at the top of the shockmount. The impedance, measured at 1000 megacycles, varied from 13.5 to 14.8 ohms and decreased with the compression of the mount. The resonant frequency of all these jumpers was in the vicinity of 3500 mc so that operation below 1000 mc is considered satisfactory.

APPENDIX XVI

THEORY OF SHIELDING

An exact solution to any shielding problem can be obtained only by solving a so-called boundary-value problem. This means that a solution to Maxwell's equations, which govern the behavior of all electromagnetic fields, must be found which satisfies the boundary conditions imposed by the metallic surfaces both of the source of the electromagnetic field and of the conductors used as shields. Boundary - value problems belong to the most difficult problems of electrical engineering, and usually an effort is made to find an approximate solution to a practical shielding problem by other means. A very fruitful approach is that of treating the problem as a boundary-value problem only for the conductors used as shields and making some kind of reasonable assumption about the field distribution on the source-side of the shield. Then the boundary conditions at the source are not necessarily satisfied, and in particular the reaction of the shield on the source is neglected, i.e., it is assumed that the source is not affected by any wave reflected from the shield. This omission is partially compensated by treating a variety of different sources since a source that is modified by the presence of a reflected wave may behave simply like some other kind of source.

The shielding problem encountered in practice may usually be formulated in the following way: Given a source of electromagnetic disturbances, usually in the form of one or several conductors carrying alternating electric currents, and a point in space, find the reduction in the electric and magnetic field intensities at the given point which is brought about by the interposition of one or several metallic sheets of specified shape and material between the source and the given point. The problem may be simplified immediately according to the statement above by replacing the source by a specified field configuration, such as would be produced, for example, by an idealized point or line source.

Even after this first simplification, the general problem is still much too complicated to allow a simple solution. If the shape of the shield is at all irregular, the analytical approach becomes all but hopeless. Only the three simplest regular shapes will be treated here, and in each case it will be assumed that the fields exhibit the same kinds of symmetries as the shields themselves. The cases to be treated are (1) that of a plane shield of infinite extent in the presence of plane waves, (2) that of a shield in the form of a circular cylinder in the presence of cylindrical waves concentric with the shield, and (3) that of a spherical shield in the presence of spherical waves concentric with the shield. While none of these three cases is exactly duplicated in practice, almost every practical case can be closely approximated by one or a combination of several of these. Therefore, the results that will be obtained here are of considerable practical importance despite their restrictions.

In any shielding problem, a valuable aid is offered by the analogy between the propagation of electromagnetic waves through a medium and the transmission of electrical energy through a transmission line. This analogy has been extensively exploited in the literature, and it is especially valuable to the engineer who is much more familiar with transmission-line theory than with propagation phenomena. It

is found that such familiar concepts as characteristic impedance, propagation constant, reflection and transmission factors, attenuation, and phase shift, all have their counterparts in the propagation of electromagnetic waves. Some of these require re-definition with a slight change in meaning, but once the new definitions have been introduced properly, all the laws of transmission-line theory can be applied directly. For example, just as the characteristic impedance of a transmission-line is defined as the ratio of voltage to current along a line on which no reflected wave is present (i.e., along a line that is either infinitely long or terminated in its characteristic impedance), so the "intrinsic impedance" of a medium is defined as the ratio of electric to magnetic field intensity in a medium in which no reflected wave is present. In general, the analogy requires that the electric field intensity in the medium be substituted for the voltage along the transmission line, and that the magnetic field intensity be substituted for the current. It should be noted that the units for electric field intensity are volts per meter or volts per inch and those for magnetic field intensity are amperes per meter or amperes per inch, so that their ratio, the intrinsic impedance, is measured in ohms just as is the characteristic impedance of a transmission line. The units of length drop out when the ratio is taken.

There is one complication in the extension of transmission-line concepts to the propagation of electromagnetic waves: The impedance is a function not only of the properties of the medium, but also of the type of wave that is being considered. It is true that the same complication arises also in transmission-line theory at high frequencies when modes of transmission other than the fundamental may be present. Then the impedances associated with the various modes will, in general, be different. But the presence of higher modes in transmission lines is a comparatively rare phenomenon in practice, and usually the term "transmission-line theory" refers to the behavior of the fundamental mode only. In the propagation of electromagnetic waves, on the other hand, it must be remembered that, at all frequencies, the impedance is a function not only of the mode, when several different modes exist, but also of the type of wave, i.e., the impedance is different for plane, cylindrical, and spherical waves.

1. Plane Waves

It is assumed that the field intensities are functions of one space coordinate only. The electric field intensity is assumed to have an x-component only and the magnetic field intensity to have only a z-component. The variation then takes place along the y-axis, which is also the direction of propagation. With these assumptions Maxwell's equations reduce to the following form:

$$\frac{\partial E_x}{\partial y} = \mu \frac{\partial H_z}{\partial t} \tag{1}$$

$$\frac{\partial H_z}{\partial y} = \sigma E_x + \varepsilon \frac{\partial E_x}{\partial t} \tag{2}$$

where E_x and H_z are the electric and magnetic field intensities, respectively, μ is the permeability of the medium, σ its conductivity, and ε its permittivity. x, y, and z are rectangular coordinates forming a right-handed system, and t is the time.

Assuming sinusoidal time variations of all field intensities at the angular frequency $\omega = 2\pi f$, where f is the frequency, the solution to Equations (1) and (2) may be written in the following form:

$$E = E_1 e^{\gamma y} + E_2 e^{-\gamma y} \tag{3}$$

$$H = \frac{E_1}{Z_o} e^{\gamma y} - \frac{E_2}{Z_o} e^{-\gamma y} \tag{4}$$

where $E = E_x e^{j\omega t}$, $H = H_z e^{j\omega t}$, $j^2 = -1$, E_1 and E_2 are arbitrary constants which must be determined from the boundary conditions, and γ and Z_o are defined as follows:

$$\gamma = \text{propagation constant of the medium} = \sqrt{j\omega\mu(\sigma + j\omega\varepsilon)} \tag{5}$$

$$Z_o = \text{intrinsic impedance of the medium} = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\varepsilon}} \tag{6}$$

Both γ and Z_o are complex quantities, in general, so that one can write:

$$\gamma = \alpha + j\beta \tag{7}$$

$$Z_o = R_o + jX_o \tag{8}$$

where α is called the attenuation constant and β the phase constant of the medium, and R_o and X_o are the resistive and reactive parts, respectively, of the intrinsic impedance.

For free space or air, $\sigma = 0$, $\mu = 4\pi \times 10^{-7}$ henries per meter, and $\varepsilon = 8.85 \times 10^{-12}$ farads per meter. With these values substituted, the intrinsic impedance of free space or air is 376.6 ohms and its propagation constant is $j\omega/c = 2\pi/\lambda$, where $c = 3 \times 10^8$ meters per second is the velocity of light in free space and λ is the wave length.

For all metals, the conductivity, σ, is much larger than the product $\omega\varepsilon$ at all frequencies now used or likely to be used for radio communication purposes. Therefore, the term $j\omega\varepsilon$ may be neglected and one obtains for metals

$$\gamma = \sqrt{\frac{\omega\mu\sigma}{2}}(1 + j) \tag{9}$$

$$Z_o = \sqrt{\frac{\omega\mu}{2\sigma}}(1 + j) \tag{10}$$

$$\alpha = \beta = \sqrt{\frac{\omega\mu\sigma}{2}} \tag{11}$$

$$R_o = X_o = \sqrt{\frac{\omega\mu}{2\sigma}} \tag{12}$$

It is seen that the intrinsic impedance of metals is extremely small in comparison with that of air and most other dielectric media. On the other hand, both the real and the imaginary parts of the propagation constant are very large in comparison with those of dielectrics, which indicates a large attenuation and a small wave length in metals.

In order to apply these results to shielding problems, the behavior of the wave at the boundary between two media must be investigated. Let it be assumed that a plane wave is propagated in air and impinges on a plane metal surface. For simplicity, choose a set of rectangular coordinates so that the boundary surface coincides with the plane $y = 0$. Let the region $y > 0$ be air and the region $y < 0$ be the metal. Let the incident wave be propagated in the direction of decreasing y, as shown in Figure XVI-1. The problem is to find a set of solutions of the type given by Equations (3) and (4) which satisfy the boundary conditions, viz. the conditions that both E and H must be continuous everywhere, including the surface $y = 0$. The transmission-line analogy of this situation is shown in Figure XVI-2: Two semi-infinite lines of different characteristic impedances are joined so that each line is terminated at one end in the characteristic impedance of the other.

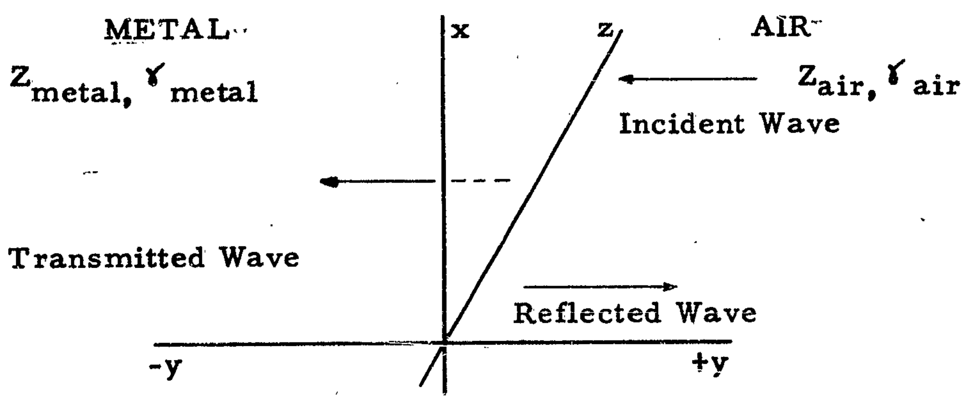


Fig. XVI-1 Plane Wave Striking Plane Boundary Surface Between Air and Metal

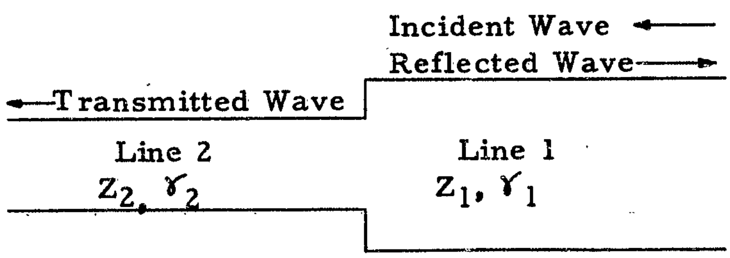


Fig. XVI-2 Transmission-Line Analogy of Plane-Wave Boundary Problem

The problem is solved by postulating the existence of a transmitted wave in the metal and a reflected wave in air traveling in the direction of increasing y, in addition to the incident wave. One must write:

$$E_{air} = E_1 e^{\gamma y} + E_2 e^{-\gamma y} \tag{13}$$

$$E_{metal} = E_1' e^{\gamma y} \tag{14}$$

$$H_{air} = \frac{E_1}{Z_{air}} e^{\gamma y} - \frac{E_2}{Z_{air}} e^{-\gamma y} \tag{15}$$

$$H_{metal} = \frac{E_1'}{Z_{metal}} e^{\gamma y} \tag{16}$$

Here E_1 and E_1 / Z_{air} are the amplitudes of the incident wave, E_2 and E_2 / Z_{air} the amplitudes of the reflected wave, and E_1' and E_1' / Z_{metal} the amplitudes of the transmitted wave. E_1, E_2, and E_1' are three constants, two of which may be evaluated from the conditions that, at $y = 0$, E_{air} must equal E_{metal} and H_{air} must equal H_{metal}. The third, which is best taken as the amplitude of the incident wave, remains arbitrary, of course. The ratio of the amplitude of the electric field intensity of the reflected wave to that of the incident wave is called the reflection factor, F_r, and the ratio of the electric field intensity of the transmitted wave to that of the incident wave is called the transmission factor, F_t. Using the boundary conditions stated, one obtains after a short computation

$$F_r : \frac{E_2}{E_1} = \frac{Z_{metal} - Z_{air}}{Z_{metal} + Z_{air}} \tag{17}$$

$$F_t = \frac{E_1'}{E_1} = \frac{2 Z_{metal}}{Z_{metal} + Z_{air}} \tag{18}$$

These expressions are identical with the defining equations for the reflection and transmission factors of a transmission line.

The quantity of interest here is 20 log F_t, which gives, in decibels, the attenuation experienced by the electric field intensity in entering the metal. It represents a reflection loss, due to the fact that the incident wave is only partially transmitted, the rest being reflected. It must be distinguished from the absorption loss within the metal, which will be discussed below.

A wave, in passing through an actual shield, experiences reflections at two boundary surfaces: Once when it enters the shield, and again when it leaves the shield. The transmission factor for the second surface may be obtained from Equation (18) by replacing Z_{metal} with Z_{air} and vice versa. The reflection loss in this case is not equal to 20 log F_t because the first medium does not extend to infinity. There will exist multiple reflections between the two surfaces of the shield, which will affect the transmission loss at the second surface. However, as will be seen presently, the absorption loss in the metal is so large that, for all practical shields, the effect of multiple reflections may be neglected.

The reflection factor was defined above as the ratio of the amplitudes of the electric field intensities. A similar factor might be defined as the ratio of the amplitudes of the magnetic field intensities. A simple calculation shows that the reflection loss for the magnetic field at the first surface is just equal to the reflection loss for the electric field at the second surface (neglecting multiple reflections), and vice versa. Thus, while the attenuation experienced by the magnetic field at any one surface is quite different from that experienced by the electric field (much smaller at the first and much larger at the second surface), the combined effect of the two surfaces is the same for both fields, as it must be in accordance with the assumption that the character of the wave does not change in passing through the shield and, therefore, that the impedances on both sides of the shield are equal.

As was shown before, in all practical cases Z_{metal} is much smaller than Z_{air} and may be neglected in the denominator of Equation (18). Using this approximation for both surfaces, one obtains for the combined reflection loss in decibels for either the electric or the magnetic field:

$$\begin{aligned}\text{Total reflection loss} &= 20 \log (Z_{air}/2\,Z_{metal}) + 20 \log (Z_{air}/2\,Z_{air}) \\ &= 20 \log (Z_{air}/4\,Z_{metal})\end{aligned} \tag{19}$$

Here the ratio is inverted as compared to Equation (18) so that the reflection loss is positive when the wave is attenuated.

In addition to the reflection loss, there is an absorption loss within the metal. Equations (14) and (16) show that both the electric and the magnetic fields within the metal contain the propagation factor $\exp \gamma y = (\exp \alpha y)(\exp j\beta y)$. The second of these factors is a phase factor that does not affect the amplitude of the fields. The first factor shows that the amplitudes decrease exponentially within the metal. If the thickness of the shield is S, the amplitudes at the second surface are smaller than those at the first surface within the metal by a factor $\exp(-\alpha S)$. The negative sign arises because within the metal y is negative. Hence, the absorption loss is

$$\text{Absorption loss} = \alpha_{metal} S \tag{20}$$

Equations (19) and (20) may be combined. After substituting numerical values

from Equations (10) and (11) and converting to practical units, one obtains for the total loss in decibels:

$$\text{Total loss} = 3.34 \sqrt{f_m \mu_r \sigma_r S} + 108.2 + 10 \log (\sigma_r / f_m \mu_r) \qquad (21)$$

where f_m is the frequency in megacycles per second, μ_r the relative magnetic permeability of the shielding material ($\mu_r = 1$ for all non-magnetic materials), σ_r the relative conductivity of the shielding material ($\sigma_r = 1$ for copper), and S the thickness of the shield in mils. For example, the total loss experienced by a plane wave in passing through a plane copper sheet 5 mils thick at 1 mc would be 124.9 decibels. In Figure XVI-3, the absorption loss, the reflection loss, and the total loss are plotted as functions of frequency for a plane copper sheet 5 mils thick. It is seen that at low frequencies the reflection loss is by far the larger, but at high frequencies the absorption loss becomes the larger. For a copper sheet 5 mils thick, the two losses are equal at about 30 mc, but for a thicker sheet, that point would occur at a lower frequency. The total loss has a minimum at about 0.3 mc, and the minimum loss is about 122 decibels.

Decreasing the conductivity decreases both the reflection and the absorption losses, so that the shielding material should always have as high a conductivity as possible. Increasing the relative permeability, i.e., using magnetic materials, increases the absorption loss but decreases the reflection loss. To obtain some idea of the effects of using magnetic materials for shielding purposes, Figure XVI-4 shows the absorption loss, the reflection loss, and the total loss plotted as a function of frequency for a material having a constant relative permeability of 1000. Since all known magnetic materials have conductivities considerably lower than copper, the relative conductivity was given a representative value of 0.1. It is seen that the total loss now reaches its minimum at about 3 kc, and the minimum loss itself is about 103 db. Thus it appears that, for the example chosen, the magnetic material is preferable at the higher frequencies, but poorer at the lower frequencies. It must be remembered, however, that the material in the example chosen is quite thin (5 mils thick). For a thicker sample, the reflection loss remains the same, but the absorption loss is increased proportionally. Since the magnetic material has the higher absorption loss, it follows that, the thicker the shield, the more advantageous does the use of magnetic materials become.

Another fact to remember is that the permeability of all magnetic materials decreases with frequency. For a typical sample having a permeability of 1000 at frequencies up to 10 mc, the permeability might decrease to 100 at 100 mc, to 10 at 1000 mc, and down to unity at 10,000 mc. Thus, the extremely high losses at the high frequencies shown in Figure XVI-4 cannot be realized in practice, but they also are not usually required.

Equation (21) was derived on the basis of a plane wave striking a plane surface normally. If the direction of propagation of the wave makes an angle with the normal to the surface other than zero degrees, the situation is more complicated. It is then necessary to treat two cases separately: that of the electric field intensity being parallel to the surface, and that of the magnetic field intensity being parallel to the surface. In the first case one speaks of a transverse electric wave; in the second

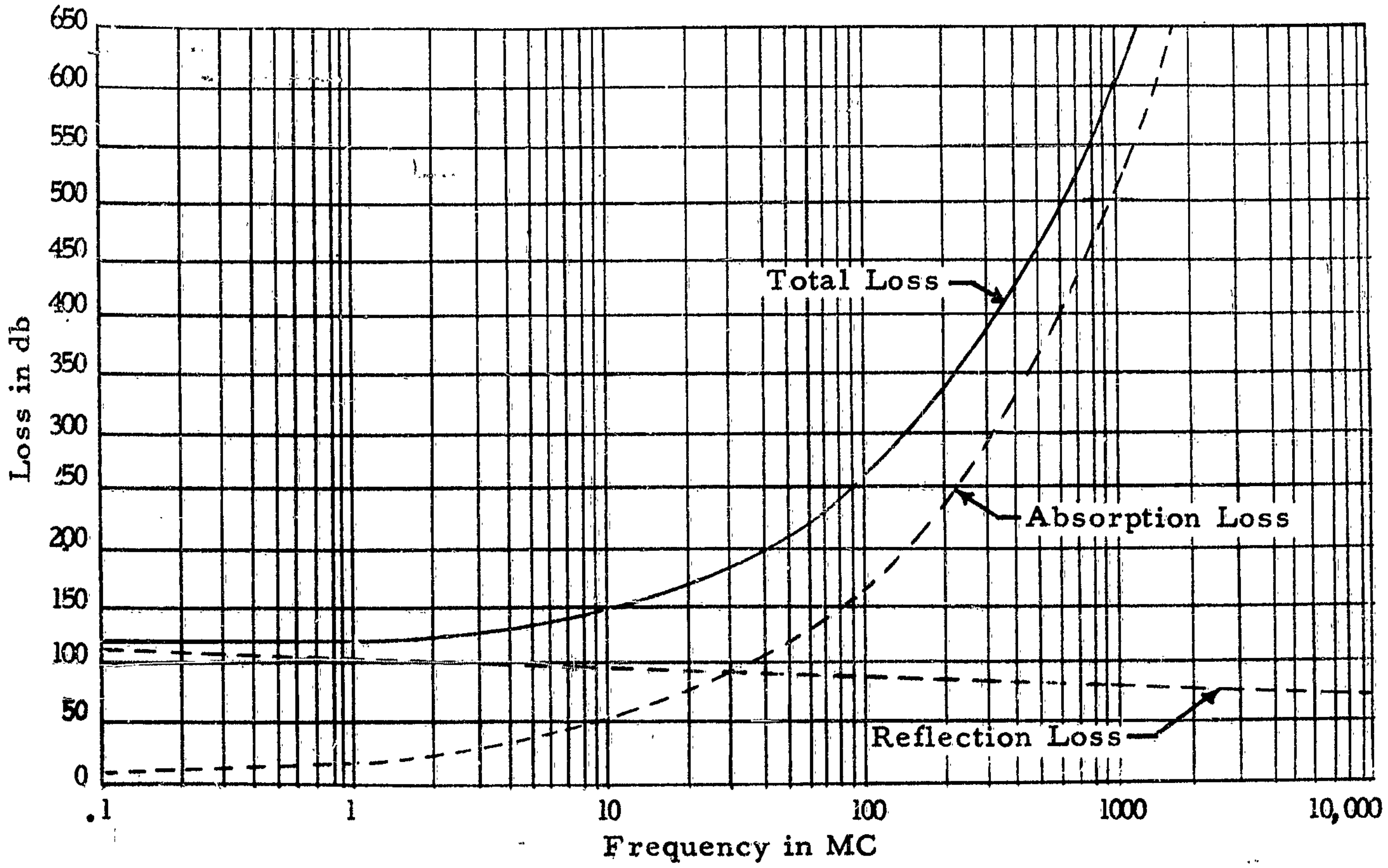


Fig. XVI-3 Absorption Reflection and Total Losses in 5 Mil Sheet of Copper

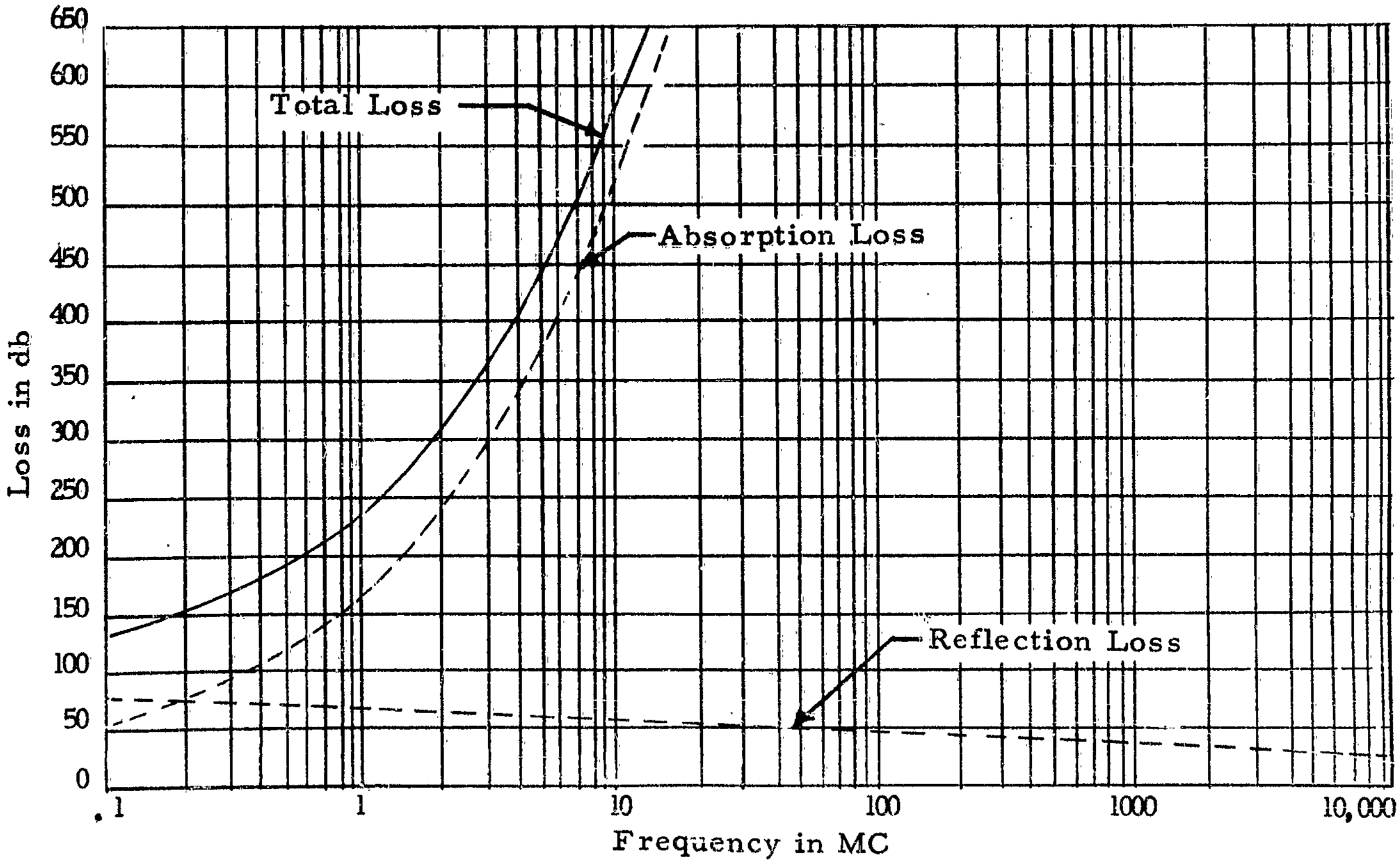


Fig. XVI-4 Absorption, Reflection, and Total Losses in
5 Mil Sheet of Magnetic Material. $\sigma r = 0.1$, $\mu_r = 1000$

case of a transverse magnetic wave. In the case of normal incidence the two cases merge into one. The analysis is simplified by the fact that, no matter at what angel a plane wave strikes a plane metal surface, the refracted wave, which is transmitted into the metal, always travels in the direction normal to the surface. Therefore, the absorption loss within the metal and the reflection loss at the second surface are independent of the angle of incidence. The detailed analysis is not carried out here, but it can be shown that the reflection loss at the first surface for oblique incidence is never less than for normal incidence, so that Equation (21) gives a total loss which is at worst too low.

Finally, it must be pointed out that Equation (21) is not valid in practical cases for either very small values of S or very small values of f. This is evident from the fact that, if one sets $S = 0$, Equation (21) still indicates a reflection loss. Actually, $S = 0$ means that there is no shield, and, therefore, there can be no loss. Also, setting $f = 0$, one finds an infinite reflection loss, while it is known that shields are very ineffective for static magnetic fields. The reasons for these discrepancies are not immediately obvious.

Since multiple reflections within the shield were neglected in the derivation of Equation (21), one might think that this omission causes the equation to cease being valid for small values of S. This, however, is not the case. Analysis shows that Equation (21) is theoretically correct, even when multiple reflections are considered, for values of S down to the order of 10^{-5} mils. This is a distance of the order of magnitude of an atomic diameter. Hence, Equation (21) is theoretically valid even for a shield consisting of a monatomic or monomolecular layer of metal. Yet, it is known that very thin shields, such as coatings of metallic paints, are sometimes very ineffective. The fallacy lies in the fact that Maxwell's equations themselves, with their macroscopic concepts of conductivity, permeability, and dielectric constant, cease to be valid in the realm of atomic or molecular phenomena. In other words, Equation (21) is valid only as long as one deals with shields that are thick enough to allow the application of concepts, such as conductivity, that require the presence of a very large number of atoms or molecules. While no definite limit can be set, it should be expected that Equation (21) will begin to break down for thicknesses less than about 0.01 or 0.001 mils.

As to the application of Equation (21) to very low frequencies, there is nothing in its derivation that would indicate that it might cease to be valid for any communication frequency. And, indeed, it does remain valid even for very low frequencies for the conditions under which it was derived: the presence of plane waves. In practice, any wave becomes approximately plane at a distance of several wave lengths from the source. At very low frequencies the wave length may be several miles, and it is usually quite impossible to even approximate a plane wave. Hence, at power frequencies and, in the limit, at zero frequency, plane waves never exist, in practice, and Equation (21) is not applicable. In addition, Maxwell's equations remain satisfied if any arbitrary constant static field is added to the solution given by Equations (3) and (4). This constant field was neglected entirely in the treatment given above. Hence, the results are not applicable to a static field except in the special case of a static field obtained from a plane wave in the limit as the frequency goes to zero, which case is never met in practice.

2. Cylindrical Waves

Practical cases involving cylindrical symmetry are encountered frequently. Out of the many possible cases, only a very small number will be investigated here. The most important examples are (1) electromagnetic waves traveling axially in a cylindrical enclosure either with an internal conductor (coaxial transmission line) or without (circular wave guide); (2) electromagnetic waves traveling radially and originating from a centrally located line current source; and (3) electromagnetic waves traveling radially and originating from a centrally located line loop source. These cases do not normally occur separately. Usually there exists a superposition of several occuring simultaneously. The first case is best attacked from the point of view of surface transfer impedance (see Appendix XI) and will not be treated here.

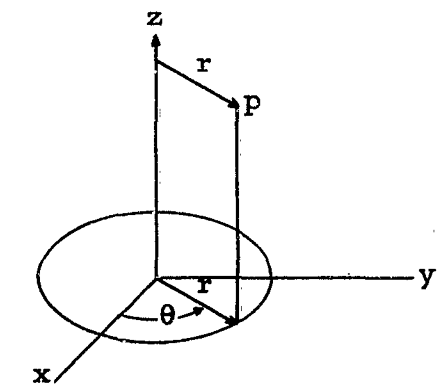


Fig. XVI-5 Cylindrical Coordinates

For the second and third case, it is assumed that the fields have circular symmetry and do not vary along the axis of the cylinder. Consider the cylindrical coordinate system shown in Figure XVI-5. It is assumed, then, that the fields are functions of r only, independent of z and θ. With these assumptions, Maxwell's equations read:

$$\frac{\partial E_z}{\partial r} = \mu \frac{\partial H_\theta}{\partial t} \tag{22}$$

$$\frac{\partial}{\partial r}(r H_\theta) = r\left(\sigma E_z + \varepsilon \frac{\partial E_z}{\partial t}\right) \tag{23}$$

$$\frac{\partial H_z}{\partial r} = -\left(\sigma E_\theta + \varepsilon \frac{\partial E_\theta}{\partial t}\right) \tag{24}$$

$$\frac{\partial}{\partial r}(r E_\theta) = -\mu r \frac{\partial H_z}{\partial t} \tag{25}$$

It is seen that this set naturally splits up into two pairs: the pair involving E_z and H_θ, Equations (22) and (23), and the pair involving E_θ and H_z, Equations (24) and (25). Each pair may be treated separately. Setting $E_\theta = H_z = 0$, one is led to

case (1) above: the fields act as if they were produced by a source in the form of an infinitely long and infinitely thin conductor at the center carrying a current that is everywhere in phase. It will be seen that such a field, in the vicinity of the source, is associated with a very low impedance, and therefore, this case will be called the "low-impedance case". If one sets $E_z = H_\theta = 0$, case (2) results: the fields act as if they were produced by an infinite number of very small, closely spaced coaxial loops carrying equal and uniform currents that are everywhere in phase. This field is associated with a high impedance near the source and will be called the "high-impedance case".

Just as in the case of plane waves, a static field may be superimposed on the dynamic fields. Such a field might be produced by a static line charge at the center and would vary as 1/r. The possible presence of this field will be neglected here.

The solution to the pair of Equations (22) and (23) may be written as follows, assuming sinusoidal time variations as before:

$$E_z = E_1 H_o (j\gamma r) + E_2 H_o' (j\gamma r) \tag{26}$$

$$H_\theta = -\frac{j}{Z_o} \left[E_1 H_1 (j\gamma r) + E_2 H_1' (j\gamma r) \right] \tag{27}$$

These equations are very similar to Equations (3) and (4) for plane waves except that the exponential functions are replaced by the functions H_o, H_o', H_1, and H_1'. These functions are the Hankel functions of first and second order and of the first and second kind. The subscripts refer to the order of the Hankel function; the unprimed symbols are Hankel functions of the first kind, and the primed symbols Hankel functions of the second kind. The Hankel functions of the first kind represent waves traveling radially inward, while the Hankel functions of the second kind represent waves traveling radially outward.

The impedance is again found as the ratio of the electric to magnetic field intensity, but a complication arises due to the fact that the impedances for the incoming and outgoing waves are not equal. One obtains from Equations (26) and (27), treating the incoming and outgoing parts separately:

$$Z_1 = j Z_o H_o (j\gamma r) / H_1 (j\gamma r) \tag{28}$$

$$Z_1' = j Z_o H_o' (j\gamma r) / H_1' (j\gamma r) \tag{29}$$

where Z_1 and Z_1' are the radial impedances for the incoming and outgoing waves, respectively, in the low-impedance case. It is seen that these impedances are functions of r.

If the same analysis is carried through for the high-impedance case, starting with Equations (24) and (25), the following set of impedances is obtained:

$$Z_2 = j Z_o H_1(j\gamma r) / H_o(j\gamma r) \tag{30}$$

$$Z_2' = j Z_o H_1'(j\gamma r) / H_o'(j\gamma r) \tag{31}$$

where Z_2 and Z_2' are the radial impedances for the incoming and outgoing waves, respectively, in the high-impedance case.

An extensive analysis of these impedances requires a detailed consideration of the behavior of the Hankel functions. This will not be carried out here. Great simplifications result if the absolute value of the argument $j\gamma r$ is either very large or very small. If it is large, i.e., at large distances from the center of the cylinder, all impedances approach the value of Z_o for plane waves. Hence, as might have been expected, cylindrical waves behave like plane waves at large distances from the center. When the absolute value of $j\gamma r$ is very small, the following approximations may be used:

$$\left.\begin{aligned} H_o(x) &\longrightarrow 1 + (2j/\pi)\ln|x| \\ H_o'(x) &\longrightarrow 1 - (2j/\pi)\ln|x| \\ H_1(x) &\longrightarrow -2j/(\pi x) \\ H_1'(x) &\longrightarrow 2j/(\pi x) \end{aligned}\right\} \text{as } x \longrightarrow 0 \tag{32}$$

With these approximations, one obtains for the four impedances in the case of small values of $j\gamma r$:

$$Z_1 = -(1/2)\, j\gamma r Z_o (\pi + 2j \ln|j\gamma r|) \tag{33}$$

$$Z_1' = -(1/2)\, j\gamma r Z_o (\pi - 2j \ln|j\gamma r|) \tag{34}$$

$$Z_2 = -\frac{2 Z_o}{j\gamma r(\pi^2 + 4\ln^2|j\gamma r|)}(\pi - 2j\ln|j\gamma r|) \tag{35}$$

$$Z_3 = -\frac{2 Z_o}{j\gamma r(\pi^2 + 4\ln^2|j\gamma r|)}(\pi + 2j\ln|j\gamma r|) \tag{36}$$

It is seen that the impedances for the incoming and outgoing waves are negative complex conjugates of each other in each case. It is also seen that, for small $j\gamma r$, the first two impedances are much smaller than Z_o while the second set of two impedances are much larger than Z_o, thus justifying the nomenclature used.

Now let there be a metallic medium filling all space outside a cylindrical region such that for r < a the medium is air and for r > a the medium is a good conductor as shown in Figure XVI-6. The surface r = a is the boundary surface between the two media, and the task at hand is to determine how much of the incident wave, originating at the center, is transmitted into the metal.

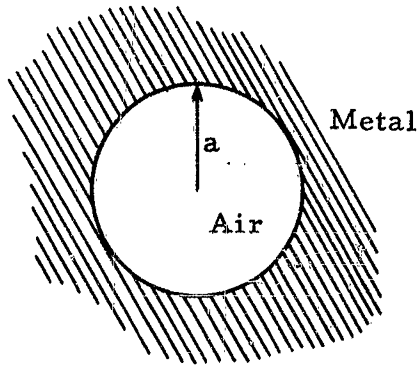


Fig. XVI-6 Cylindrical Boundary Surface Between Air and Metal

As in the case of plane waves, the presence of an incident and a reflected wave must be postulated on the inside together with a transmitted wave within the metal. To obtain a general expression for the transmission factor, taking into consideration the fact that the impedances for outgoing and incoming waves need not be equal, the following equations may be written based on the continuity of the electric and magnetic fields across the boundary surface:

$$E'_{metal} = E'_{air} + E_{air} \tag{37}$$

$$\frac{E'_{metal}}{Z'_{metal}} = \frac{E'_{air}}{Z'_{air}} - \frac{E_{air}}{Z_{air}} \tag{38}$$

where the primed quantities refer to the outgoing waves and the unprimed ones to the incoming waves. In addition, the subscripts 1 or 2 must be added depending on whether the low-impedance case or the high-impedance case is dealt with.

The above equations may be solved for the transmission factor, E'_{metal}/E'_{air}:

$$F_t = \frac{Z'_{metal}}{Z'_{metal} + Z_{air}} \left(1 + \frac{Z_{air}}{Z'_{air}}\right) \tag{39}$$

It is seen that this reduces to Equation (18) when the primed and unprimed impedances are equal.

When the quantity $|\gamma|$ is computed for a metal from Equation (9), one finds $|\gamma| = 545 \sqrt{f_m \mu_r \sigma_r}$ per inch. Hence the absolute value of the quantity $j\gamma r$ will be much larger than unity in most practical applications. Therefore, within the metal cylindrical waves behave like plane waves. In particular, the absorption loss within the metal is the same as that obtained for plane waves, Equation (20), and the impedance of the metal to cylindrical waves is the same as for plane waves. An exception

to this might occur only when the quantity $r\sqrt{f_m \mu_r \sigma_r}$ (r in inches) is less than, say, 0.1, and even then the assumption of plane waves will be a fair approximation down to $r\sqrt{f_m \mu_r \sigma_r} = 0.01$, approximately. The exceptional case in which this might not be true will not be treated here.

In evaluating the four impedances for air from Equations (33) to (36), it is found that Z_1 and Z_1' (low-impedance case) are very small while Z_2 and Z_2' (high-impedance case) are very large as compared with Z_o in the region of validity of these equations. A more detailed computation of these quantities is hardly justified since the physical conditions on which the derivation of these equations was based are highly idealized. In practice, the presence of an inner conductor imposes additional boundary conditions, which were neglected here. However, the following qualitative conclusion drawn from the form of Equation (39) remains valid: In the high-impedance case the reflection loss is at least as great as with plane waves; but in the low-impedance case the reflection loss may be very small and cannot be relied upon for effective shielding action.

The practical conclusions for cylindrical shields may be summarized as follows:

(a) Within the shielding metal, the cylindrical waves behave like plane waves. The absorption loss may be computed using Equation (20) as for plane waves.

(b) If the electric field near the source at the center is predominantly axial, the radial impedance is low, and very little shielding action may be expected due to reflection.

(c) If the electric field near the source at the center is predominantly concentric, the radial impedance is high, and the shielding action due to reflection is at least as good as that for plane waves.

(d) In the absence of detailed information about the fields near the source, it is best to assume that a substantial low-impedance component is present, and to proceed on the assumption that shielding action is due to absorption only.

3. Spherical Waves

Practical cases involving spherical symmetry are very rare in shielding problems encountered in aircraft. If the same approach is taken in this case as was taken for cylindrical waves, similar results are obtained except that the resulting functions are spherical Hankel functions, or Hankel functions of half-integral order, instead of the cylindrical Hankel functions H_o, H_1, H_o', and H_1'. This analysis, however, is not carried out here.

A somewhat different approach was taken in Paragraph 1.6.4, where Equations (14) to (25) give the fields in the vicinity of a point source. No boundary conditions were imposed there, and the fields described are those of waves traveling outward from a specified source. Nevertheless, even there a low-impedance case and a high-impedance case may be distinguished. If, in these equations, the ratio of the

electric to the magnetic field intensity is taken, it is found that near the source this ratio is proportional to the distance r for the magnetic dipole, while it is proportional to 1/r for the electric dipole. Hence, the first case, where the electric field is entirely concentric, is a low-impedance case, while the second case, where the electric field near the source is entirely axial, i.e., along a diameter, is a high-impedance case.

If the radius of the sphere is very much larger than a wave length, the waves behave exactly like plane waves. If the radius of the sphere is very much smaller than a wave length, the conclusions (a) to (d) at the end of the section on cylindrical waves remain essentially valid. The intermediate case, where resonant excitation of the spherical enclosure may occur, is again too complicated to be treated here. The main reason why spherical waves are mentioned here is that the conclusions just mentioned remain valid qualitatively when a point source is shielded by an enclosure of other than spherical shape, as for example when a spark gap, acting like a point source, is enclosed by a rectangular shielding box. Here, an exact solution of the boundary problem involved is practically impossible. Yet, it may be said that the shielding effectiveness of the box is at least as great as that computed on the basis of the absorption loss for plane waves, and may be considerably increased by the reflection loss if the source is of the high-impedance type.

APPENDIX XVII

DECIBEL EQUIVALENTS TO CURRENT, VOLTAGE, AND POWER RATIOS

The word Decibel is frequently used to express the ratio of any two dimensionally equivalent quantities in convenient numerical terms. It is simply a logarithm of a ratio to the base ten. In practice, the engineer often refers to a quantity as so many decibels "up" or "down" with respect to another quantity. This means that, if quantity a is "up" a given number of decibels on quantity b, the ratio of a/b is greater than unity, or if the a is down on b, the ratio a/b is less than unity. Sometimes +db is used to express the fact that the ratio is greater than one and -db for ratios less than one. Since the engineer always can use the reciprocal of any ratio to obtain the same result when expressed in decibels, it is never necessary to use a conversion table showing ratios less than unity.

The number of decibels, N_{db}, corresponding to a power ratio P_1/P_2 or P_2/P_1 is defined as

$$\pm N_{db} = 10 \log_{10}(P_1/P_2) \text{ or } 10 \log_{10}(P_2/P_1) \qquad (1)$$

If $P_1/P_2 > 1$, P_1 is up on (larger than) P_2 and the sign of N_{db} may be taken as +. However, if $P_1/P_2 < 1$, that is $P_2/P_1 > 1$, P_1 is down on (less than) P_2 and the sign is -.

In the case of direct currents, power is the product of the current I and the voltage E, that is P = IE. Since the resistance R is defined as E/I, the power may be taken as $P = E^2/R$ or $P = I^2R$. Thus $P_1/P_2 = (E_1/E_2)^2 \times (R_2/R_1)$ or $P_1/P_2 = (I_1/I_2)^2 \times (R_1/R_2)$. This means that provided that the currents I_1 and I_2 are flowing into the same resistance ($R_1 = R_2$) the power ratio may be expressed as

$$\pm N_{db} = 10 \log_{10}(I_1/I_2)^2 \qquad (2)$$

$$= 20 \log_{10}(I_1/I_2) \qquad (3)$$

or also as

$$\pm N_{db} = 20 \log_{10}(E_1/E_2) \qquad (4)$$

In the case of alternating currents, the product IE must be multiplied by the power factor in order to obtain the power. But if now the currents I_1 and I_2 are flowing into the same impedance ($Z_1 = Z_2$), the power factor will cancel when the ratio P_1/P_2 is formed, and the expressions given above remain correct provided that I_1, I_2, E_1, and E_2 are interpreted as the absolute values of these quantities. It should be noted that the first expression, involving the currents, remains correct even for different impedances provided only that the resistances are equal. But the second expression, involving the voltages, is correct only if both the resistances and the reactances are equal.

In case the designer desires to obtain the current, voltage, or power ratio when the number of decibels is known, the relations are

$$|I_1/I_2| = 10^{\pm N_{db}/20} \qquad (5)$$

$$|V_1/V_2| = 10^{\pm N_{db}/20} \tag{6}$$

or

$$P_1/P_2 = 10^{\pm N_{db}/10} \tag{7}$$

keeping in mind, as before, that the number of decibels corresponding to any value of the ratios is the same as that for the reciprocal, but the sign is different.

The table given in this appendix, Figure XVII-B, may be entered to obtain the ratios of currents, voltages, and power corresponding to decibel numbers from 0.1 to 20.0. This range of decibels covers ratios of current and voltage from 1.012 to 10.00 and ratios of power from 1.023 to 100.

The designer of radio interference suppression equipment will not frequently have occasion to deal with ratios or decibel values smaller than those shown in the table, Figure XVII-B. In those cases, however, the designer may easily compute any of the values lying below the range of the table from the equations given above or the approximate formula, $N_{db} = 8.5(|I_1/I_2| - 1)$, with an error less than 0.001 db.

Decibels	Current of Voltage Ratio $\lvert I_1/I_2 \rvert$ or $\lvert E_1/E_2 \rvert$	Power Ratio P_1/P_2
10	3.162	10
20	10.000	10^2
30	3.162×10	10^3
40	10^2	10^4
50	3.162×10^2	10^5
60	10^3	10^6
70	3.162×10^3	10^7
80	10^4	10^8
90	3.162×10^4	10^9
100	10^5	10^{10}
110	3.162×10^5	10^{11}
120	10^6	10^{12}
130	3.162×10^6	10^{13}
140	10^7	10^{14}
150	3.162×10^7	10^{15}
160	10^8	10^{16}

Fig. XVII-A Decibel Values Corresponding to Power Ratios and Absolute Values of Current or Voltage Ratios. (For reciprocal ratios use negative sign for decibels. For negative decibel values take reciprocal ratios.)

Usually radio interference problems demand consideration of larger values of attenuation. Figure XVII-A is helpful in determining the ratios for larger values of decibels in 10 db steps from 10 to 160 db and may be used in conjunction with Figure XVII-B to quickly obtain intermediate values as in the following examples:

(a) To determine current ratio corresponding to 43.5 db, enter Figure A and

note that 40 db is equivalent to a current ratio of 100. Entering Figure B shows that 3.6 db is equivalent to a ratio of 1.514. Then to obtain the ratio for 43.6 db, multiply 1.514 x 100 which gives 151.4.

(b) To find the voltage ratio for 56.7 db, enter Figure A and find 3.162×10^2 opposite 50 db, multiply this by 2.163 from Figure B opposite 6.7 db and obtain 683.9. Or to find the corresponding power ratio enter the same tables in the power column and get 100,000 x 4.677 or 467,700.

(c) Conversely, to find the number of decibels corresponding to a current ratio of 2427 divide by 1000 = 10^3, enter Figure B opposite 2.427 and find 7.7 decibels, enter Figure A opposite 10^3, and find 60 decibels opposite 10^3, then add to obtain 67.7 db.

Deci-bels	$\lvert I_1/I_2\rvert$ $\lvert E_1/E_2\rvert$	P_1/P_2	Deci-bels	$\lvert I_1/I_2\rvert$ $\lvert E_1/E_2\rvert$	P_1/P_2	Deci-bels	$\lvert I_1/I_2\rvert$ $\lvert E_1/E_2\rvert$	P_1/P_2	Deci-bels	$\lvert I_1/I_2\rvert$ $\lvert E_1/E_2\rvert$	P_1/P_2
0.1	1.012	1.023	5.1	1.799	3.236	10.1	3.199	10.23	15.1	5.689	32.36
0.2	1.023	1.047	5.2	1.820	3.311	10.2	3.236	10.47	15.2	5.754	33.11
0.3	1.035	1.072	5.3	1.841	3.388	10.3	3.273	10.72	15.3	5.821	33.88
0.4	1.047	1.096	5.4	1.862	3.467	10.4	3.311	10.96	15.4	5.888	34.67
0.5	1.059	1.122	5.5	1.884	3.548	10.5	3.350	11.22	15.5	5.957	35.48
0.6	1.072	1.148	5.6	1.906	3.631	10.6	3.388	11.48	15.6	6.026	36.31
0.7	1.084	1.175	5.7	1.928	3.715	10.7	3.428	11.75	15.7	6.095	37.15
0.8	1.096	1.202	5.8	1.950	3.802	10.8	3.467	12.02	15.8	6.166	38.02
0.9	1.109	1.230	5.9	1.972	3.891	10.9	3.508	12.30	15.9	6.237	38.90
1.0	1.122	1.259	6.0	1.995	3.981	11.0	3.548	12.59	16.0	6.310	39.81
1.1	1.135	1.288	6.1	2.018	4.074	11.1	3.589	12.88	16.1	6.383	40.74
1.2	1.148	1.318	6.2	2.042	4.169	11.2	3.631	13.18	16.2	6.457	41.69
1.3	1.162	1.349	6.3	2.065	4.266	11.3	3.673	13.49	16.3	6.531	42.66
1.4	1.175	1.380	6.4	2.089	4.365	11.4	3.715	13.80	16.4	6.607	43.65
1.5	1.189	1.412	6.5	2.114	4.467	11.5	3.758	14.13	16.5	6.683	44.67
1.6	1.202	1.445	6.6	2.138	4.571	11.6	3.802	14.45	16.6	6.761	45.71
1.7	1.216	1.479	6.7	2.163	4.677	11.7	3.846	14.79	16.7	6.839	46.77
1.8	1.230	1.514	6.8	2.188	4.786	11.8	3.891	15.14	16.8	6.918	47.86
1.9	1.245	1.549	6.9	2.213	4.898	11.9	3.936	15.49	16.9	6.998	48.98
2.0	1.259	1.585	7.0	2.239	5.012	12.0	3.981	15.85	17.0	7.080	50.12
2.1	1.274	1.622	7.1	2.265	5.128	12.1	4.027	16.22	17.1	7.161	51.29
2.2	1.288	1.660	7.2	2.291	5.248	12.2	4.074	16.60	17.2	7.244	52.48
2.3	1.303	1.698	7.3	2.317	5.370	12.3	4.121	16.98	17.3	7.328	53.70
2.4	1.318	1.738	7.4	2.344	5.495	12.4	4.169	17.38	17.4	7.413	54.95
2.5	1.334	1.778	7.5	2.371	5.623	12.5	4.217	17.78	17.5	7.499	56.23
2.6	1.349	1.820	7.6	2.399	5.754	12.6	4.266	18.20	17.6	7.586	57.54
2.7	1.365	1.862	7.7	2.427	5.888	12.7	4.315	18.62	17.7	7.674	58.88
2.8	1.380	1.906	7.8	2.455	6.026	12.8	4.365	19.05	17.8	7.763	60.26
2.9	1.396	1.950	7.9	2.483	6.166	12.9	4.416	19.50	17.9	7.852	61.66
3.0	1.413	1.995	8.0	2.512	6.310	13.0	4.467	19.95	18.0	7.943	63.10
3.1	1.429	2.042	8.1	2.541	6.457	13.1	4.519	20.42	18.1	8.035	64.57
3.2	1.445	2.089	8.2	2.570	6.607	13.2	4.571	20.89	18.2	8.128	66.07
3.3	1.462	2.138	8.3	2.600	6.761	13.3	4.624	21.38	18.3	8.222	67.61
3.4	1.479	2.188	8.4	2.630	6.918	13.4	4.677	21.88	18.4	8.318	69.18
3.5	1.496	2.239	8.5	2.661	7.080	13.5	4.732	22.39	18.5	8.414	70.79
3.6	1.514	2.291	8.6	2.692	7.244	13.6	4.786	22.91	18.6	8.511	72.44
3.7	1.531	2.344	8.7	2.723	7.413	13.7	4.842	23.44	18.7	8.610	74.13
3.8	1.549	2.399	8.8	2.754	7.586	13.8	4.898	23.99	18.8	8.710	75.86
3.9	1.567	2.455	8.9	2.786	7.763	13.9	4.955	24.55	18.9	8.811	77.62
4.0	1.585	2.512	9.0	2.818	7.943	14.0	5.012	25.12	19.0	8.913	79.43
4.1	1.603	2.570	9.1	2.851	8.128	14.1	5.070	25.70	19.1	9.016	81.28
4.2	1.622	2.630	9.2	2.884	8.318	14.2	5.128	26.30	19.2	9.120	83.18
4.3	1.641	2.692	9.3	2.917	8.511	14.3	5.188	26.92	19.3	9.226	85.11
4.4	1.660	2.754	9.4	2.951	8.710	14.4	5.248	27.54	19.4	9.333	87.10
4.5	1.679	2.818	9.5	2.985	8.913	14.5	5.309	28.18	19.5	9.441	89.13
4.6	1.698	2.884	9.6	3.020	9.120	14.6	5.370	28.84	19.6	9.550	91.20
4.7	1.718	2.951	9.7	3.055	9.333	14.7	5.433	29.51	19.7	9.661	93.33
4.8	1.738	3.020	9.8	3.090	9.550	14.8	5.495	30.20	19.8	9.772	95.50
4.9	1.758	3.090	9.9	3.126	9.772	14.9	5.559	30.90	19.9	9.886	97.72
5.0	1.778	3.162	10.0	3.162	10.000	15.0	5.623	31.62	20.0	10.000	100.00

Fig. XVII-B Decibel values corresponding to Power Ratios and absolute value of Current and Voltage ratios. (Use negative db values for reciprocal ratios or reciprocal ratios for negative db values)

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